Before the
FEDERAL COMMUNICATIONS COMMISSION
Washington, D.C. 20554

In the Matter of	)	
	)	
Revision of Part 15 of the Commission's	)	
Rules Regarding Ultra-Wideband	)	ET Docket No. 98-153
Transmission Systems	)	

COMMENTS

1. The Commission has adopted a Notice of Inquiry ("NOI") in the above-captioned matter,[1] requesting comments on a number of issues related to the possible operation of ultra-wideband ("UWB") devices under Part 15 of its Rules. These devices would emit sequences of very short-duration (*e.g.,* 0.1 to 10 nanoseconds) pulses, and would have extremely wide (possibly 10 GHz or more) emission bandwidths. The Commission is specifically requesting comments on the potential of UWB devices to interfere with other radio devices and services, and how UWB devices might be managed from a regulatory perspective (*i.e.,* emission limits, measurement procedures, and usage restrictions) to prevent harmful interference.

2. The Wireless Information Networks Forum ("WINForum") hereby offers its comments in response to the NOI. WINForum's membership includes manufacturers of unlicensed devices operating under Part 15 in the 1.9 GHz Unlicensed PCS ("UPCS") band, the Industrial, Scientific, and Medical ("ISM") bands, and the recently-adopted Unlicensed National Information Infrastructure ("U-NII") bands near 5 GHz.

3. WINForum has been an active participant in Commission proceedings related to the operation of unlicensed devices. WINForum's WINTech committee was instrumental in developing the spectral "etiquette" for the UPCS allocation, which is now Subpart D of Part 15. WINForum's WINTest committee, in cooperation with the Commission's Office of Engineering and Technology ("OET") and committee C63 of the American National Standards Institute ("ANSI") developed detailed measurement

[1] FCC 98-208, adopted August 20, 1998; released September 1, 1998.

procedures, adopted by ANSI as a American National Standard (ANSI C63.17-1998) in March of this year, for verifying compliance of UPCS devices with Subpart D. Most recently, WINForum and Apple Computer filed Petitions for Rule Making that led to the U-NII Rules, and WINForum's 5 GHz Sharing Rules Development Committee ("SRDC") worked with the Commission and other interested parties throughout the entire cycle of the U-NII proceeding to develop and refine the U-NII Rules. The SRDC continues to take an active interest in regulatory issues related to the operation of U-NII devices.

4. WINForum has interest in the instant proceeding for two reasons. First, WINForum believes that regulations should be developed for UWB devices to prevent or minimize interference that could disrupt the operation of wireless devices, including those in the UPCS, ISM, and U-NII bands. Second, WINForum believes it can contribute to the proceeding by virtue of its expertise in test and measurement techniques. WINForum therefore offers these comments.

5. WINForum first notes that currently, potential interference sources typically have their energy confined to a narrow frequency range. A victim receiver therefore tends to be affected only by devices with energy concentrated near the receiver's center frequency. However, with UWB devices, this is not the case because of the large UWB bandwidth. A victim receiver may be subject to interference from all UWB devices within interfering range.

6. To answer the questions posed in the NOI regarding interference from UWB devices, emission limits, and measurement procedures, it is necessary to understand the effect of a UWB transmission on a receiver which has a bandwidth that is narrow relative to the bandwidth of the UWB emission. Such a narrowband receiver could represent either a measuring instrument, such as a spectrum analyzer, or a "victim" receiver suffering UWB interference. To develop this understanding, WINForum has conducted detailed analyses, documented in Attachments 1 and 2.

7. Attachment 1 develops a mathematical model for describing a UWB signal and its effect on receiver that is narrowband relative to the UWB pulse. It is shown that if the receiver bandwidth exceeds the UWB pulse rate, the effective interference power varies

as the square of the receiver bandwidth. If the receiver bandwidth is less than the pulse rate and the pulse rate is constant, the interference power varies as the square of the pulse rate, independent of the receiver bandwidth. It is demonstrated with analysis and simulation results that under some conditions, this conclusion also applies to a UWB transmission that uses pulse-position modulation (PPM) to convey information. If the interval between UWB pulses varies randomly, the interference appears to be broadband noise with an average power that varies as the average pulse rate multiplied by the receiver bandwidth, as long as the average pulse rate is much greater than the receiver bandwidth.

8. It also is shown in Attachment 1 that in all of the above cases, the interference power is proportional to the energy spectral density of the individual pulse, and does not depend directly on the peak pulse power or total pulse energy. The "pulse densensitization factor," which relates the peak power observed from a narrowband measurement to the peak power of a wideband pulse, therefore is not relevant. Moreover, the pulse densensitization calculation requires knowledge of the pulse duration, which often will not be measurable using a conventional instrument such as a spectrum analyzer.

9. Attachment 2 examines the efficacy of the existing radiated emission limits codified in §15.109 and the associated measurement procedures in §15.35. This is done by analyzing the effect on potential victim receivers of five different emission types which meet the existing limits. It is shown that the effect varies widely with emission type and that the current rules do not properly account for this variation.

10. Based on the results in Attachments 1 and 2, it is clear that the existing limits do not provide adequate protection to some types of receivers against interference from some types of emissions, including those from UWB devices. WINForum believes that, based on these analyses, a power spectral density ("PSD") limit should be imposed for at least some types of devices, including UWB devices. Potential victim devices with bandwidths covering a fairly wide range (e.g., 3 kHz to 30 MHz) should be protected by such a limit.

11. In the case of UWB devices, a single-bandwidth measurement to test for compliance with a PSD limit would be inadequate, because the effective PSD of the interference depends on the pulse repetition rate and the receiver bandwidth. However, the overview at the beginning of Attachment 1 outlines some potential measurement procedures that could be used to calculate the PSD for a wide range of bandwidths based on a limited number of measurements. It appears that the required measurements could be made with a spectrum analyzer. In the case of devices which vary the inter-pulse interval (*e.g.*, with PPM), information from the manufacturer providing parameters such as the average pulse rate and the modulation scheme may be necessary to correctly interpret measurement results and verify compliance with the PSD limit over the specified range of bandwidths.

12. In summary, WINForum believes that the existing Part 15 limits for unwanted emissions, and the associated measurement procedures, are not adequate for controlling emissions from UWB devices. Moreover, they cannot adequately regulate interference from disparate transmission formats in a consistent manner, as shown in Attachment 2 to these comments. It appears that in most cases, a PSD limit would provide the most consistent protection. However, compliance with peak limits is sometimes easier to measure, and peak limits will suffice in some cases. In the case of UWB devices, measurements must be made in a way that allows the PSD to be calculated over a range of bandwidths, because for a UWB device, the PSD itself depends on the bandwidth of the victim receiver.

13. In the NOI, the Commission raised a question about the effect of aggregate interference from multiple UWB devices. Although WINForum has not yet addressed that issue, it clearly is very important because of the potential of UWB devices to cause interference over a wide frequency range, and the potential for widespread proliferation. WINForum therefore plans to investigate the aggregate interference issue.

14. WINForum believes that the material provided in the Attachments to these Comments will be useful to the Commission in its consideration of the appropriate regulations for UWB transmissions, and is prepared to assist the Commission in evaluating potential emission limits and measurement procedures.

Respectfully submitted,

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ATTACHMENT 1

ANALYSIS OF ULTRA-WIDEBAND TRANSMISSIONS

Measurement and Interference Issues

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December 7, 1998

OVERVIEW AND CONCLUSIONS

Purpose

The FCC has released a Notice of Inquiry (NOI) requesting comments pertaining to potential emission limits, measurement procedures, and usage restrictions for ultra wideband (UWB) devices. These devices transmit sequences of extremely short pulses (on the order of 0.1 to 2 nanoseconds in duration), resulting in emission bandwidths which can exceed 10 GHz. Many of the questions in the NOI concern the potential of UWB devices to interfere with other radio devices and services, and how UWB devices might be managed from a regulatory perspective to prevent harmful interference.

This paper develops a mathematical model for describing of the effect of UWB transmissions on receivers that are narrowband compared to the UWB bandwidth. Most, if not all, potential victim receivers as well as conventional measurement instruments fall into this category. Therefore, the results derived here can be used to assess interference potential as well as to evaluate measurement procedures.

Summary of Results

The results can be reduced to a fairly simple set of relationships, which can be summarized as follows. The UWB signal is assumed to consist of a very short-duration pulse which is repeated at some rate R_p pulses per second. If the pulse repetition rate is constant (no modulation) then the spectrum of the pulse sequence consists of spectral lines at frequencies that are multiples of the pulse rate. Of interest is the effect of the UWB signal on a filter of bandwidth B_h, which is assumed to be much less than the bandwidth of the pulse. That effect is described here in terms of the filter output power, which represents interference or a measured power level (e.g., on a spectrum analyzer).

The filter will have some center frequency f_0, and the energy spectral density of the pulse at that frequency is denoted as $\Phi(f_0)$. The filter output depends on $\Phi(f_0)$ as well as the pulse repetition discipline and the filter bandwidth. There are four cases which are explored here:

[1] If the pulse is repeated at regular intervals (no variation in the time between pulses), then the spectrum of the UWB signal consists of spectral lines at frequencies that are multiples of R_p. The power in the spectral component at frequency kR_p is $\Phi(kR_p)R_p^2$. This is the power output of a filter which is narrow enough to resolve the spectral lines; i.e., $B_h < R_p$. Stated in terms of the time domain, the filter response time exceeds the inter-pulse interval, so the filter output is due to the combined effect of multiple UWB pulses.

[2] If the inter-pulse interval is varied randomly and $B_h << \overline{R}_P$ (where $\overline{R}_P$ is the average repetition rate), the filter output will have a probability distribution approaching that of Gaussian noise, with average power $\Phi(f_0)B_h\overline{R}_P$. As the filter bandwidth B_h approaches the pulse rate, the filter output will become less noise-like. For a filter bandwidth that exceeds the pulse rate, case [4] below applies.

[3] If pulse-position modulation (PPM) is used with an average pulse rate $\overline{R}_P$, and $B_h < \overline{R}_P$, there will in general be spectral lines of varying strengths at some frequencies that are integer multiples of $\overline{R}_P$, and the strongest lines will have power $\Phi(k\overline{R}_P)\overline{R}_P^2$. The positions and strengths of the spectral lines depends on the pulse-position deviation relative to the nominal inter-pulse interval $1/\overline{R}_P$.

[4] If the filter bandwidth exceeds the pulse rate (regardless of the pulse repetition discipline), then the filter responds to each pulse individually, or in frequency-domain terms, the filter bandwidth spans multiple spectral lines and cannot resolve them. In this case, the filter power output is $\Phi(f_0)B_h^2$. This is the average power output over the filter response time. The absolute peak envelope power is about 3 dB higher, due to peaking of the filter impulse response. The quantity $\Phi(f_0)B_h^2$ seems to be the appropriate measure of interference potential, since if it is large enough it will cause a symbol error in the victim receiver. The effect of these symbol errors will depend on the nature of the system supported by the victim receiver as well as the rate at which the errors occur, but in some cases, periodic bit errors could effectively cause link failure.

<u>Implications for Emission Limits and Measurement Procedures</u>

Note that in all cases, the filter power output depends on the energy spectral density of the pulse, not on the peak power or total energy in the pulse. This suggests that a limit on peak pulse power is not important, with respect to controlling interference. Also, compliance with such a limit would be difficult to verify with a spectrum analyzer. The “pulse densensitization factor” for calculating the peak power in a wideband pulse from a spectrum analyzer measurement is based on a rectangular pulse shape, and requires knowledge of the pulse duration. The duration of a UWB pulse cannot be measured by a spectrum analyzer. In some cases, it can be inferred from the width of spectral sidelobes, but this approach cannot be reliably used for UWB devices because of uncertainty about the pulse shape and spectrum shaping by the UWB antenna. Thus, the "pulse desensitization" calculation does not seem to be useful or practical for evaluating the interference potential of UWB devices.

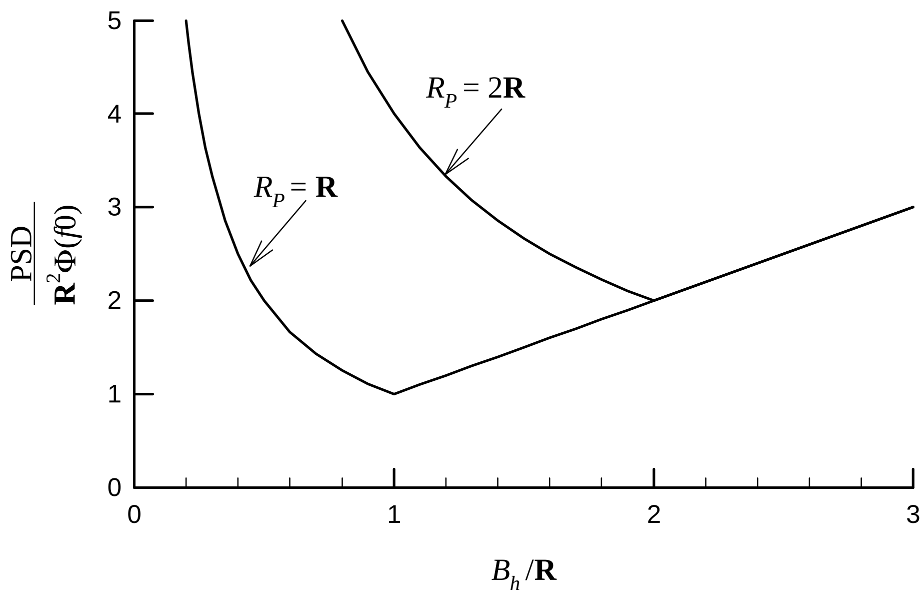
One reasonable measure of the effect of interference is the amount by which it raises the effective noise floor of the victim receiver. That increase is determined by the ratio of the interference power (p) to the victim bandwidth, or p/B_h, which is effectively the power spectral density (PSD) of the interference. From the results summarized above, the filter power output and the effective power spectral density are as follows.

Constant pulse rate or PPM with $B_h < R_p$: $p = \Phi(f_0)R_p^2$ $PSD = \Phi(f_0)R_p^2/B_h$

Random pulse interval with $B_h < R_p$: $\overline{p} = \Phi(f_0)B_h R_p$ $PSD = \Phi(f_0)R_p$

Any repetition discipline with $B_h > R_p$: $p = \Phi(f_0)B_h^2$ $PSD = \Phi(f_0)B_h$

The figure below shows the PSD for two different UWB transmissions that use the same pulse, but one has a repetition rate **R** and the other has rate 2**R**.



As can be seen, for a given pulse rate, the minimum PSD occurs when $B_h = R_p$. This means that if a single measurement bandwidth is used, UWB devices with pulse rates at or near that bandwidth will tend to be favored.

For a bandwidth of 2**R** or more, the PSD is the same for these two devices, so if a measurement were made with that bandwidth, their interference potential would appear to be the same. However, for victim devices with bandwidths of **R** or less, the UWB device with the higher pulse rate causes 6 dB more interference.

If the energy spectral density of the higher-rate system is reduced by 6 dB, then for bandwidths less than **R**, the PSD is the same for the two systems. However, for bandwidths greater than 2**R**, the higher-rate system causes 6 dB less interference than the lower-rate system.

The point is that a single-bandwidth emission limit is inadequate for controlling the interference potential of UWB devices, because the effective power spectral density of the interference caused by UWB emissions will depend not only on the energy spectral density of the pulse and the repetition rate, but also on the bandwidth of the victim receiver relative to the pulse repetition rate.

It is apparent that emission limits for UWB devices will need to be structured to provide protection for potential victim devices with a wide range of bandwidths. For example, bandwidths of several kHz or less can be used for long-distance communications such as CW or low-rate data links. At the other extreme, C-band point-to-point system use bandwidths in the range of 20-30 MHz with high-efficiency modulation techniques such as 64-ary quadrature amplitude modulation (QAM), which require a high carrier-to-noise ratio to achieve the desired reliability. One possible way to structure regulations would be to establish a PSD limit (watts/Hz), which would apply over a range of bandwidths $B_{\min} \le B_h \le B_{\max}$. As an example, $B_{\min}$ and $B_{\max}$ might be 3 kHz and 30 MHz, respectively.

To verify compliance, measurements could be made directly using a spectrum analyzer, up to the maximum resolution bandwidth of the instrument (typically 1-2 MHz for a conventional analyzer). Beyond that, the interference power could in some cases be calculated from the data measured with lower bandwidths. For example, if spectral lines can be resolved using some relatively narrow resolution bandwidth, and the pulse rate of the UWB device is constant, then the rate is equal to the spectral line separation and the maximum energy spectral density $\Phi_{\max}$ is easily calculated. The power output of a narrowband filter can be computed for any desired bandwidth from the formulas above.

One potential pitfall to this approach is that if PPM is used, spectral lines may be evident but at a separation that is some multiple of the average pulse rate. The spectrum may look similar to that of a constant-rate emission, which could lead to an error in calculating the repetition rate and therefore the energy spectral density. As an example, if PPM is used with a pulse position shift of ±25% about the pulse “clock”, depending on whether “0” or “1” is transmitted, the spectrum consists of lines separated by $2\overline{R}_p$, each with power $\Phi(f_0)\overline{R}_p^2$. This spectrum could easily be interpreted as the spectrum of a constant-rate sequence of rate $2\overline{R}_p$ and energy spectral density $\Phi(f_0)/4$. The calculation of the interference potential to a wideband victim receiver would therefore be 6 dB too low.

If spectral lines cannot be resolved using even a very narrow resolution bandwidth, then it is likely that either: (1) the repetition rate is very low (less than the narrowest available resolution bandwidth); or (2) the inter-pulse interval is varying in some random or pseudorandom way. In the first case, the peak measured power will increase quadratically as the resolution bandwidth is increased, per the equations above, regardless of whether the pulse repetition rate is constant. Another way to test for case (1) is to view the signal in zero-span mode, so that the filter response to individual pulses is apparent if $B_h > \overline{R}_p$. If case (1) applies, then $\Phi_{\max}$ can be computed based on measured power and the resolution bandwidth used to make the measurement, and the PSD can be calculated for any desired bandwidth. If case (2) applies, then the resolution bandwidth can be increased until responses to individual pulses can be seen separately in zero-span mode. At that point, the measured power is 3 dB above $\Phi(f_0)B_h^2$, so the power output for any desired bandwidth greater than B_h can be calculated. If case (2) applies and the individual pulses cannot be resolved, even with the maximum resolution bandwidth of the instrument, then the peak-to-average ratio can be used to calculate the average pulse rate. The average power is $\overline{p} = \Phi(f_0)B_h\overline{R}_p$ and the peak is $p_{\text{peak}} = \Phi(f_0)\overline{R}_p^2$, assuming $B_h < \overline{R}_p$. Both can easily be measured with a spectrum analyzer, using a narrow video filter for the average, and a wide video filter (3 times the resolution bandwidth) with "max hold" for the peak. The ratio is $p_{\text{peak}} \big/ \overline{p} = \overline{R}_p \big/ B_h$, and B_h is about 1.13 times the nominal resolution bandwidth. Knowing $\overline{R}_p$, $\Phi_{\max}$ can be calculated and the PSD for any desired bandwidth can be calculated.

Conclusions

Overall, it appears that for most cases, the parameters needed to verify UWB device compliance with a PSD limit can be either measured directly with a spectrum analyzer, or calculated based on measurements, although PPM has the potential for a spectrum that can create some confusion about the average pulse rate. Manufacturers therefore should be required to provide data about the pulse rate and modulation technique for such systems.

Based on the result derived in this paper, the following specific questions posed in the NOI can be answered as shown in italics following the questions.

- Are the existing general emission limits sufficient to protect other users of the spectrum, especially radio operations in the restricted bands, from harmful interference? ***No; more general limits need to be established, based on the maximum power spectral density over a range of bandwidths.***
- Should different limits be applied to UWB systems? ***Yes, as discussed above.***
- Should we specify a different standard for UWB devices based on spectral power density? ***Yes. In fact, consideration should be given as to whether power density***

limits should apply to other types of devices as well. Should these standards be designed to ensure that the emissions appear to be broadband noise? ***Power spectral density limits themselves will tend to encourage designers to make the emissions affect a narrowband receiver in the same way as broadband noise, rather than creating discrete spectral lines. However, a signal that appears to be broadband noise on a spectrum analyzer may appear as a series of discrete interfering pulses to a wideband (e.g., 20 to 30 MHz) receiver, such as are used in C-band fixed microwave systems. The effect on such wideband receivers would be greater than would be predicted from conventional measurements using a 1-MHz filter.***

- Should a limit on the total peak level apply to UWB devices? ***No. The peak power output of a UWB device does not directly impact the effective interference seen by a narrowband receiver. It is the energy spectral density of the pulse and the pulse repetition rate that are important.***
- Is a pulse desensitization correction factor appropriate for measuring emissions from a UWB device? ***No. This correction factor relates peak power measured on a spectrum analyzer to the peak pulse power, and requires knowledge of the pulse duration. It also is based on a rectangular pulse.*** Should any modifications be made to this measurement procedure for UWB devices? ***Potential measurement procedures are outlined above.***
- Would another measurement procedure that does not apply a pulse desensitization factor be more appropriate for determining the interference potential of a UWB device? ***Yes. A potential approach is outlined above.***
- Are the measurement detector functions and bandwidths appropriate for UWB devices? ***No. A single-bandwidth measurement is inadequate, because the measured PSD can vary as a function of the measurement bandwidth.*** Should these standards be modified, and if so, how? ***A PSD limit should be established, and measurements and calculations should be performed as outlined above to ensure compliance with that limit over a specified range of bandwidths.***
- Are there any other changes to the measurement procedures that should be applied to UWB devices? ***Yes, as discussed above.***

1. INTRODUCTION

The FCC has released a Notice of Inquiry (NOI) regarding ultra-wideband (UWB) transmission systems, which transmit sequences of very short-duration pulses (e.g., 100 picoseconds to 2 nanoseconds), resulting is extremely wide bandwidths (on the order of 10 GHz). The FCC is requesting comments on a number of questions in an effort to determine whether UWB devices should be allowed to operate on an unlicensed basis under Part 15, and if so, what regulations and measurement procedures should be used to prevent them from causing harmful interference to receivers of existing systems.

This analysis quantifies the effect of a UWB transmission on a receiver that is narrowband relative to the UWB transmission, which is necessary in order to understand both the measurement of UWB transmissions using a spectrum analyzer, and the effect of interference from UWB transmissions on other devices. The material developed here is intended to help answer a number of specific questions asked in the NOI related to interference, emission limits, and measurement procedures.

The approach is to develop a mathematical model for the response of a narrowband filter to a UWB transmission, and to relate the power (peak and/or average) of the filter response to the characteristics of the UWB transmission. Based on that model, answers can be developed to some of the questions in the NOI.

2. NOTATION

The following conventions will be used here

$q(t)$ a baseband pulse waveform which is non-zero for $-t/2 \le t \le t/2$ only.

$g(t)$ a single transmitted pulse. With some UWB devices, $g(t) = q(t)$; that is, the baseband pulse is applied directly to the antenna terminals. If $g(t)$ is a pulse-modulated carrier of frequency f_c, then $g(t) = q(t)\cos(2pf_c t + f)$.

$b(t) = \sum_{k=-\infty}^{\infty} q(t - kT)$ a baseband pulse train with repetition interval T

$x(t) = \sum_{k=-\infty}^{\infty} g(t - kT)$ a transmitted pulse train

$h_b(t)$ baseband-equivalent filter impulse response

$h(t) = 2h_b(t)\cos 2pf_0 t$ impulse response of filter with center frequency f_0, if $h_b(t)$ is real. In general, $h(t) = 2\,\mathrm{Re}\{h_b(t)e^{j2pf_0 t}\}$ where $\mathrm{Re}\{\cdot\}$ denotes the real part of the argument.

Unless otherwise indicated, it will be assumed that all time-domain waveforms and filter transfer functions are real (have no imaginary components). Filter impulse responses are causal; that is $h(t) = 0$ for $t < 0$.

3. FOURIER ANALYSIS

3.1 Spectrum, Energy, and Bandwidth of a Baseband Pulse

Consider a baseband pulse waveform $q(t)$, which is non-zero for $-t/2 \le t \le t/2$ only. Its Fourier transform is

$$Q(f) = \int_{-t/2}^{t/2} q(t) e^{-j2pft} dt \quad \text{volts/Hz} \tag{1}$$

The energy spectrum is $|Q(f)|^2$ joules/Hz, and the total energy in the pulse is:

$$E_q = \int_{-t/2}^{t/2} |q(t)|^2 dt = \int_{-\infty}^{\infty} |Q(f)|^2 df \tag{2}$$

The bandwidth B_q is defined here such that the total energy in a rectangular energy spectrum of bandwidth B_q and magnitude $|Q(0)|^2$ is E_q. Hence,

$$B_q = \frac{E_q}{|Q(0)|^2} \tag{3}$$

Now assume that $q(t)$ is repeated continuously at an interval *T*, forming a periodic baseband signal $b(t)$, which can be represented as a Fourier series:

$$b(t) = \sum_{k=-\infty}^{\infty} c_k e^{j2pkt/T} \tag{4}$$

with

$$c_k = \frac{1}{T} \int_{-t/2}^{t/2} q(t) e^{-j2pkt/T} dt = \frac{1}{T} Q\left(\frac{k}{T}\right) \tag{5}$$

Clearly, if $q(t)$ is real, $Q(f) = Q^*(-f)$, and $c_k = c^*_{-k}$; that is c_k and c_{-k} are complex conjugates.

3.2 Example: Rectangular Pulse

If $q(t)$ is a unipolar rectangular pulse with amplitude *A*, then $Q(f) = A\int_{-t/2}^{t/2} e^{-j2pft}dt$
$= A\dfrac{\sin pft}{pf}$ and $c_k = A\dfrac{t}{T}\dfrac{\sin kp\ t/T}{kp\ t/T}$, and L'Hopital's rule gives the dc component as $c_0 = At/T$. The bandwidth of the pulse is $1/t$.

Clearly, the spectrum is even; that is, $Q(f) = Q(-f)$, and the one-sided series for the pulse train is

$$b(t) = \sum_{k=0}^{\infty} a_k \cos 2pk\, t/T \tag{6}$$

where $a_k = c_k + c_{-k} = 2c_k$ for $k > 0$, giving

$$\begin{aligned} a_k &= 2A\frac{t}{T}\frac{\sin kp\ t/T}{kp\ t/T} \quad k > 0 \\ a_0 &= A\frac{t}{T} \end{aligned} \tag{7}$$

The power in each spectral line is

$$\begin{aligned} P_k &= \frac{a_k^2}{2} = 2A^2\left(\frac{t}{T}\right)^2\left(\frac{\sin kp\ t/T}{kp\ t/T}\right)^2 \quad k > 0 \\ P_0 &= A^2\left(\frac{t}{T}\right)^2 \end{aligned} \tag{8}$$

The peak power in the baseband pulse is $P_{PK} = A^2$ and the average power is $\overline{P} = P_{PK}\ t/T$. The spectral component power therefore can be expressed as

$$\begin{aligned} P_k &= \frac{a_k^2}{2} = 2P_{PK}\left(\frac{t}{T}\right)^2\left(\frac{\sin kp\ t/T}{kp\ t/T}\right)^2 \quad k > 0 \\ P_0 &= P_{PK}\left(\frac{t}{T}\right)^2 \end{aligned} \tag{9}$$

Hence, the power in each component varies as the square of the duty cycle, for a given peak power.

3.3 Modulation

If $q(t)$ amplitude-modulates a carrier of frequency f_c, the transmitted signal is $g(t) = q(t)\cos 2pf_c t$, and its Fourier transform is:

$$G(f) = \frac{1}{2}\left[Q(f - f_c) + Q(f + f_c)\right]. \tag{10}$$

If $g(t) = q(t)\sin 2pf_c t$, then

$$G(f) = \frac{j}{2}\left[Q(f + f_c) - Q(f - f_c)\right] \tag{11}$$

The energy spectra of (10) and (11) are the same as long as $B_q << f_c$.

If the pulse train $b(t)$ amplitude-modulates a carrier of frequency f_c, the resulting RF signal is

$$\begin{aligned} x(t) = b(t)\cos 2pf_c t &= \sum_{k=0}^{\infty} a_k \cos 2pk\, t/T \cos 2pf_c t \\ &= \frac{1}{2}\sum_{k=0}^{\infty} a_k \left[\cos 2p\left(f_c - \frac{k}{T}\right)t + \cos 2p\left(f_c + \frac{k}{T}\right)t\right] \end{aligned} \tag{12}$$

The peak envelope power (PEP) is $P_{PE} = A^2/2$. For $k > 0$, the power in each of the two spectral lines at frequencies $f_c \pm k/T$ is $a_k^2/8$. Therefore,

$$\begin{aligned} P_k &= \frac{a_k^2}{4} = 2P_{PE}\left(\frac{t}{T}\right)^2\left(\frac{\sin kp\ t/T}{kp\ t/T}\right)^2 \quad k > 0 \\ P_0 &= P_{PE}\left(\frac{t}{T}\right)^2 \end{aligned} \tag{13}$$

3.4 Example: Bipolar Sinusoidal Pulse

A UWB device may apply a "baseband" pulse directly to the antenna terminals rather than amplitude-modulating a carrier with a baseband pulse. Since the antenna frequency response will filter out low frequencies, it would be preferable to select a pulse waveform with little or no energy at low frequencies. A transmitted pulse $g(t)$ will have no energy at dc if $\int_{-\infty}^{\infty} g(t)dt = 0$. A simple example which meets this condition is the pulse

$$\begin{aligned} g(t) &= A \sin 2p\, t/t\,, \quad -t/2 \le t \le t/2 \\ &= 0\,, \quad |t| > t/2 \end{aligned}\,, \tag{14}$$

which is simply a single cycle of a sine wave of frequency $1/t$. Therefore, $g(t)$ is equivalent to a carrier of frequency $1/t$, amplitude-modulated by the rectangular pulse $q(t)$ of the previous example. Since $Q(f) = A\dfrac{\sin pft}{pf}$, (11) gives:

$$G(f) = \frac{jAt}{2}\left[\frac{\sin p(ft+1)}{p(ft+1)} - \frac{\sin p(ft-1)}{p(ft-1)}\right] \tag{15}$$

Figures 1 and 2 show $g(t)$ and the energy spectrum $|G(f)|^2$.

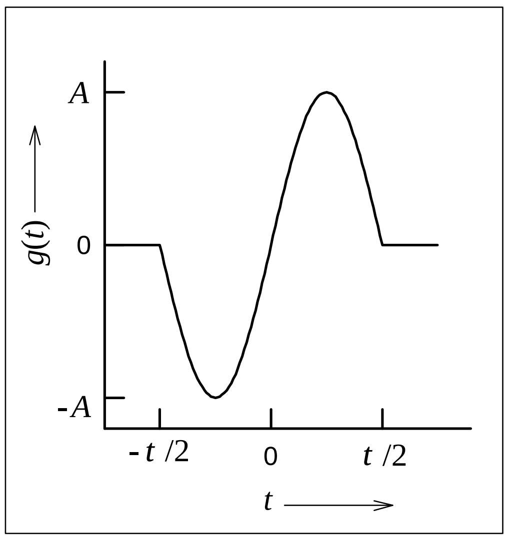


Figure 1: *Single-cycle bipolar sinusoidal pulse.*

If the pulse duration t is 1 nanosecond, then the bandwidth is on the order of 1 GHz. While this particular pulse has sidelobes with nulls separated by $1/t$ Hz, the sidelobes may be filtered out by the antenna frequency selectivity so that only the main lobe remains.

4. RESPONSE OF A NARROWBAND FILTER TO A UWB SIGNAL

This section explores the effect of UWB signals on devices that are narrowband relative to the UWB emission bandwidth. Understanding the response of a narrowband filter to a UWB signal is necessary for both the analysis of interference and assessment of measurement procedures.

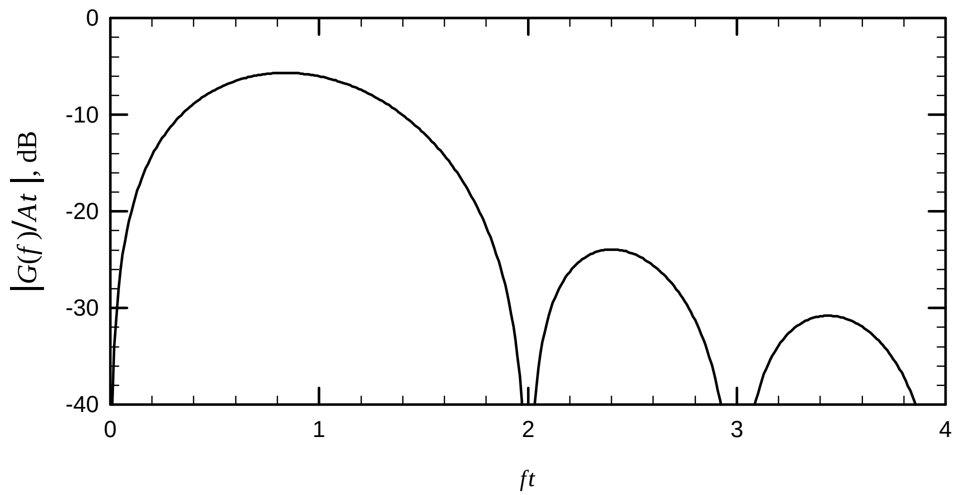


Figure 2: *Energy spectrum of single-cycle bipolar pulse.*

4.1 Filter Frequency Response, Bandwidth, and Impulse Response

Let $H(f)$ represent the frequency response of a spectrum analyzer resolution filter, or of a victim receiver, which typically is determined by the intermediate frequency (IF) section. $H(f)$ is related to the baseband-equivalent frequency response $H_b(f)$ by

$$H(f) = H_b(f - f_0) + H_b^*(-f - f_0) \tag{16}$$

where f_0 is the filter center frequency.

If $h_b(t)$ is the impulse response corresponding to $H_b(f)$, the impulse response of the bandpass filter is

$$h(t) = 2\,\mathrm{Re}\left\{h_b(t)e^{j2\pi f_0 t}\right\}. \tag{17}$$

The bandwidth of the filter is

$$B_h = \frac{\int_{-\infty}^{\infty} |H_b(f)|^2 df}{|H_b(0)|^2} \tag{18}$$

The baseband-equivalent response of an *n*-pole filter (such as a spectrum analyzer resolution filter) is discussed in Annex A. The bandpass response is:

$$H(f) = \frac{1}{\left[1 + j2\pi(f - f_0)/\alpha\right]^n} + \frac{1}{\left[1 + j2\pi(f + f_0)/\alpha\right]^n}. \tag{19}$$

In this case, $h_b(t)$ is real (see Annex A), so:

$$h(t) = 2h_b(t)\cos 2\pi f_0 t\ . \tag{20}$$

4.2 Response to a Single UWB Pulse

Let $g(t)$ represent a single UWB pulse with bandwidth B_g. It is assumed here that $B_g >> B_h$, so that $G(f)$ is approximately constant over the bandwidth of $H(f)$. The Fourier transform of the filter output $y(t)$ is:

$$Y(f) = G(f)H(f) \cong G(f_0)H_b(f - f_0) + G(-f_0)H_b^*(-f - f_0) \tag{21}$$

Since $g(t)$ is real, $G(f) = G^*(-f)$; that is, $G(f)$ is conjugate-symmetric.

Letting $G(f_0) = |G(f_0)|e^{j\phi}$, (21) becomes:

$$Y(f) = |G(f_0)|\left[H_b(f - f_0)e^{j\phi} + H_b^*(-f - f_0)e^{-j\phi}\right] \tag{22}$$

so

$$y(t) = 2|G(f_0)|\,\mathrm{Re}\left\{h_b(t)e^{j2\pi f_0 t + \phi}\right\} = 2|G(f_0)||h_b(t)|\cos[2\pi f_0 t + \phi + \theta(t)] \tag{23}$$

where $\theta(t)$ represents any time-phase variation contributed by $h_b(t)$; i.e., $h_b(t) = |h_b(t)|e^{j\theta(t)}$. If $H_b(f)$ is conjugate-symmetric, then $h_b(t)$ is real and $\theta(t) = 0$. In any event, $h(t)$ is a bandpass filter with $B_h << f_0$, so $d\theta(t)/dt << 2\pi f_0$; that is, any frequency modulation in the filter impulse response will be very small relative to f_0.

The filter output is effectively an amplitude-modulated pulse with frequency f_0 and envelope $2|G(f_0)||h_b(t)|$. The envelope power is

$$p_y(t) = 2|G(f_0)|^2|h_b(t)|^2 \tag{24}$$

The worst case clearly occurs when $|G(f)|$ has its maximum value at f_0. In that case, the output power can be related to the total energy of $g(t)$ and its bandwidth. The total energy in the pulse $g(t)$ is

$$E_g = \int_{-\infty}^{\infty}|G(f)|^2 df = \int_{-\infty}^{\infty}|g(t)|^2 dt\ . \tag{25}$$

If $|G(f)|$ has its maximum value at f_0, then the bandwidth of $g(t)$ is by definition:

$$B_g = \frac{E_g}{2|G(f_0)|^2} \tag{26}$$

Hence,

$$p_y(t) = \frac{E_g}{B_g} |h_b(t)|^2 \tag{27}$$

If $g(t)$ is a carrier of frequency f_0 that is amplitude-modulated by the rectangular pulse of width τ discussed earlier, then the peak envelope power is $p_g = E_g / t$, and $B_g = 1/t$. Letting $B_{imp} = h_{\max}(t)$ represent the “impulse bandwidth”, the peak envelope power out of the filter is

$$p_{y\,\text{peak}} = p_g \left(B_{imp} t\right)^2 \tag{28}$$

The factor $\left(B_{imp} t\right)^2$ is the “pulse desensitization” factor specified for pulsed-RF measurements in HP application note 150-2. It allows the peak pulse power to be calculated from a spectrum analyzer measurement. Note, however, that it does not apply to an arbitrary pulse shape. Also, even for a rectangular pulse, use of the desensitization factor requires knowledge of the pulse duration t. For a rectangular pulse, t is the inverse of the null-to-null bandwidth of a spectral sidelobe, so t is known if multiple lobes of the pulse spectrum are visible. However, with a UWB transmission, the antenna may filter out the sidelobes, making the pulse duration difficult to determine, even for a carrier that is amplitude-modulated by a short rectangular pulse.

The conclusion is that in general, the pulse desensitization factor will not be very useful for characterizing UWB devices. As is apparent (at least for a single pulse), the peak pulse power *per se* is not the determining factor with respect to interference potential. It is the maximum energy spectral density $|G(f)|^2_{\max}$ (joules/Hz) that is important. That is, the maximum peak envelope power output of the filter is:

$$p_{y\,\max\,\text{peak}}(t) = 2|G(f)|^2_{\max} B_{imp}{}^2 \tag{29}$$

As shown in Annex A, for an *n*-pole filter (a typical spectrum analyzer resolution filter is 4-pole), $B_{imp} \cong 1.4 B_h$, or about 1.6 times the nominal resolution bandwidth (the 3-dB bandwidth).

4.3 Response to a Periodic Pulse Sequence

If the pulse $g(t)$ is repeated at an interval T, the resulting periodic signal is

$$x(t) = \sum_{n=-\infty}^{\infty} g(t - nT) = \sum_{k=-\infty}^{\infty} c_k e^{j2\pi kt/T} \tag{30}$$

where the second sum is the Fourier series. From (5), $c_k = \frac{1}{T} G\left(\frac{k}{T}\right)$. Since

$$G(f) = G^*(-f), \tag{31}$$

$$x(t) = \frac{1}{T}\left[G(0) + 2\sum_{k=1}^{\infty} \operatorname{Re}\left\{G\left(\frac{k}{T}\right)e^{j2\pi kt/T}\right\}\right] \tag{32}$$

With $G(f) = |G(f)|e^{j\phi(f)}$,

$$x(t) = \frac{1}{T}\left\{G(0) + 2\sum_{k=1}^{\infty}\left|G\left(\frac{k}{T}\right)\right|\cos\left[2\pi k\,t/T + \phi\left(\frac{k}{T}\right)\right]\right\} \tag{33}$$

The envelope power at frequency k/T is

$$P(k/T) = \frac{2}{T^2}\left|G\left(\frac{k}{T}\right)\right|^2, \qquad k > 0 \quad . \tag{34}$$

Again, the worst case occurs when the maximum of $G(f)$ occurs at the filter center frequency f_0. In that case, assuming $B_h << \frac{1}{T}$, the envelope power of the filter output is constant and is

$$p_y(f_0) = \frac{2}{T^2}|G(f_0)|^2|H(f_0)|^2 \tag{35}$$

for $f_0 = n/T$ with n an integer.

In other words, the filter power output is due to a single spectral component of the pulse sequence $X(f)$.

If the pulse repetition frequency (PRF) $1/T$ is low compared to the filter bandwidth B_h, then the filter cannot resolve the spectral lines of $X(f)$. In this case, the filter response time is short compared to the inter-pulse interval, and the filter responds to each interfering pulse individually and the output is as in (24). This output is repeated at an interval *T*, and the envelope power output is:

$$p_y(t) = 2\left|G(f_0)\right|^2 \sum_{n=-\infty}^{\infty} \left|h_b(t-nT)\right|^2 \qquad (36)$$

The average envelope power output is

$$\overline{p}_y = \frac{2\left|G(f_0)\right|^2}{T} \int_0^{\infty} \left|h_b(t)\right|^2 dt = \frac{2\left|G(f_0)\right|^2}{T} B_h \left|H_b(0)\right|^2 \qquad (37)$$

and the peak envelope power output is

$$p_{y\,\mathrm{max\,peak}} = 2\left|G(f_0)\right|^2 \left|h_{b\,\mathrm{max}}(t)\right|^2 = 2\left|G(f_0)\right|^2 (kB_h)^2 \left|H_b(0)\right|^2 \qquad (38)$$

where kB_h is the “impulse bandwidth”. As discussed in Annex A, $k \approx 1.44$ for *n*-pole filters.

From an interference perspective, it is the energy of the interference over the duration of the symbol (of the desired signal) that is important. Assuming matched-filter detection, this energy is $2\left|G(f_0)\right|^2 \int_0^{\infty} \left|h_b(t)\right|^2 dt = 2\left|G(f_0)\right|^2 B_h \left|H_b(0)\right|^2$. The effective symbol-average interference power therefore is $2\left|G(f_0)\right|^2 B_h^2$, assuming the symbol rate is B_h and that $\left|H_b(0)\right| = 1$. Therefore, the interference power for $R_p << B_h$ will be taken as:

$$p_y = 2\left|G(f_0)\right|^2 B_h^2, \quad R_p << B_h \qquad (39)$$

This is about 3 dB below the peak, so peak measurements can be made with a spectrum analyzer and adjusted. It should be kept in mind that B_h is about 1.13 times the nominal resolution bandwith (the 3-dB bandwidth of the resolution filter).

4.4 Response to a Random Pulse Sequence

If the inter-pulse interval is random, then $x(t)$ is a random process and so is $y(t)$, which can be written as:

$$y(t) = 2\left|G(f_0)\right| \sum_{n=-\infty}^{\infty} h_b(t-T_n) \cos\left[2pf_0(t-T_n) + f\right] \qquad (40)$$

where T_n is the time at which the n^{th} pulse arrives $(T_n < T_{n+1})$, and it has been assumed that $h_b(t)$ is real. Without loss of generality f can be taken as 0. Letting $b_n = 2pf_0T_n$, (40) becomes

$$\begin{aligned} y(t) &= 2|G(f_0)|\sum_{n=-\infty}^{\infty} h_b(t-T_n)[\cos b_n \cos 2pf_0t + \sin b_n \sin 2pf_0t] \\ &= 2|G(f_0)|[C_I \cos 2pf_0t + C_Q \sin 2pf_0t] \end{aligned} \tag{41}$$

where

$$\begin{aligned} C_I &= \sum_{n=-\infty}^{\infty} h_b(t-T_n)\cos b_n \\ C_Q &= \sum_{n=-\infty}^{\infty} h_b(t-T_n)\sin b_n \end{aligned} \tag{42}$$

If $\overline{R}_P$ is the average pulse rate, then for any given value of t there will be roughly $\overline{R}_P/B_h$ pulses which contribute significantly to $y(t)$. If the $\{T_n\}$ are random, the $\{b_n\}$ (mod 2π) will tend to be uniformly-distributed between 0 and 2π, so as $\overline{R}_P/B_h$ increases, the distribution of C_I and C_Q will approach zero-mean Gaussian (Central Limit theorem). Their variances are:

$$\begin{aligned} E[C_I^2] &= E\left\{\left[\sum_{n=-\infty}^{\infty} h_b(t-T_n)\cos b_n\right]^2\right\} \\ E[C_Q^2] &= E\left\{\left[\sum_{n=-\infty}^{\infty} h_b(t-T_n)\sin b_n\right]^2\right\} \end{aligned} \tag{43}$$

Since $E[\cos b_n \cos b_m] = E[\sin b_n \sin b_m] = 0, n \neq m$,

$$E[C_I^2] = E[C_Q^2] = \frac{1}{2}\sum_{n=-\infty}^{\infty} h_b^2(t-T_n) \cong \frac{\overline{R}_P}{2}\int_0^{\infty} h_b^2(t)dt = \frac{\overline{R}_P}{2} B_h|H_b(0)|^2 \tag{44}$$

Hence, for $\overline{R}_P >> B_h$, $y(t)$ is essentially a bandpass Gaussian process with average power

$$E[y^2(t)] = 4|G(f_0)|^2 \cdot \frac{1}{2}\left(E[C_I^2] + E[C_Q^2]\right) = 2|G(f_0)|^2 \overline{R}_P B_h|H_b(0)|^2, \tag{45}$$

which is consistent with (37). The Monte Carlo simulation discussed in Annex B has been used to verify (45) and the Gaussian distribution of C_I and C_Q.

The quantity $2|G(f_0)|^2 \overline{R}_P$ represents the power spectral density of the process $x(t)$. There are no spectral lines because the $\{T_n\}$ have been assumed random. In reality, there may be constraints on the pulse positioning which will result in a spectral lines. As an example, the next subsection gives an analysis and simulation results for the case of pulse position modulation.

4.5 Response to a Pulse Position Modulated Sequence

One obvious way to transmit information on a UWB signal is to use pulse-position modulation (PPM). It is assumed here that simple binary PPM is used with an average pulse interval of *T*. The k^{th} pulse is transmitted at time $T(k+d_k)$, where $d_k = \pm d_{\max}$, depending on whether a "1" or a "0" is transmitted, with $d_{\max} < 0.5$. The Monte Carlo simulation discussed in Annex B was used to determine the average power output of a narrowband filter with bandwidth $B_h = 1/20T$ and center frequency f_0 for various values of $d_{\max}$. Figures 3, 4, and 5 show the normalized filter power output for $d_{\max} = 0.1$, 0.25, and $\sqrt{2}/10$, respectively. These graphs correspond to the display of a spectrum analyzer with the resolution bandwidth set to 1/20 the average pulse repetition rate. Spectral lines are evident in all three cases.

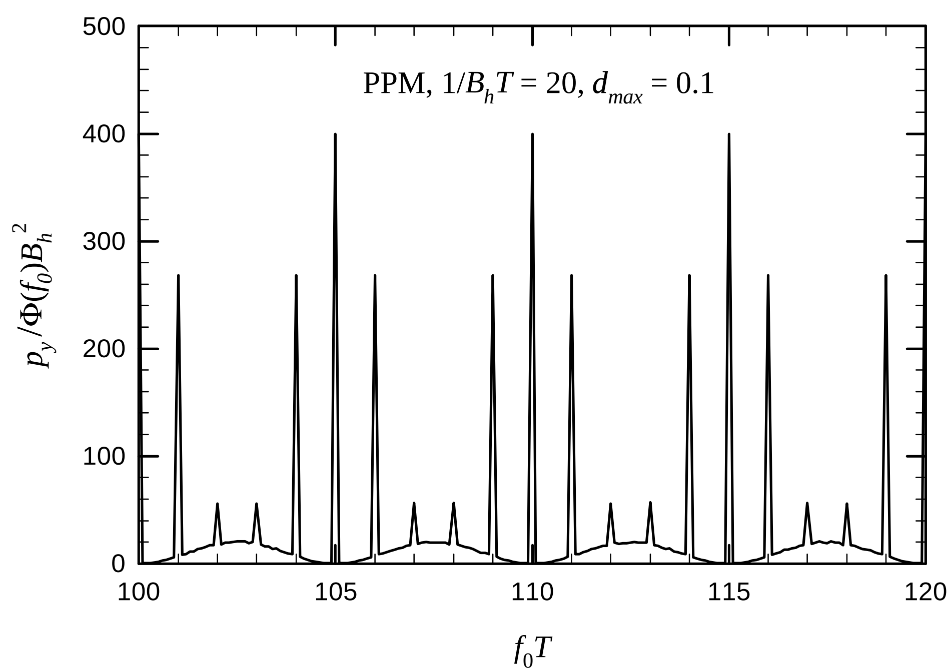


Figure 3

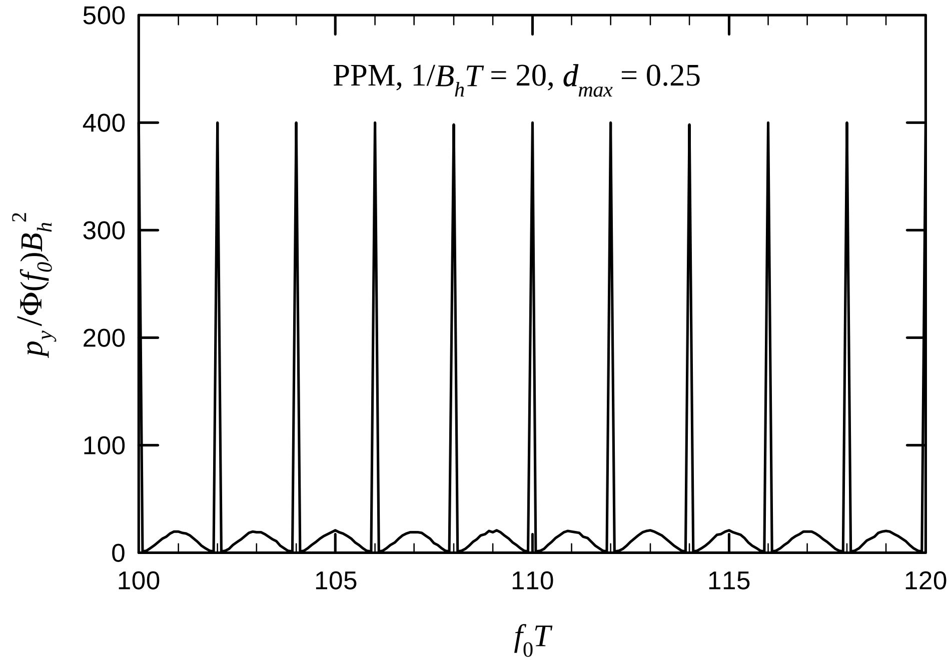


Figure 4

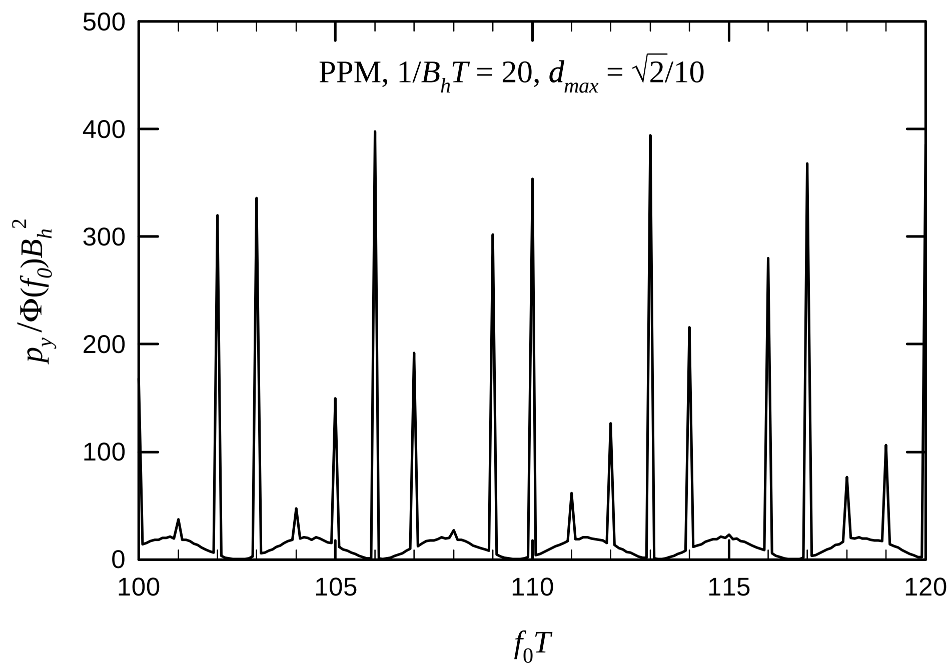


Figure 5

An expression for the magnitude of the spectral lines can be developed as follows by recognizing that when $f_0 T$ is an integer, the filter responses to all "1" pulses add constructively, and the same is true for the responses to all "0" pulses. Hence, if there are N_1 "1" pulses and N_0 "0" pulses contributing to the filter output at a given time, the normalized filter output voltage (within an arbitrary phase offset) can be written as the phasor sum:

$$\begin{aligned}\frac{y(t)}{2|G(f_0)|B_h} &= N_1 \cos 2\pi f_0 t + N_0 \cos 2\pi f_0 (t + 2\delta_{\max} T) \\ &= \cos 2\pi f_0 t (N_1 + N_0 \cos 4\pi f_0 \delta_{\max} T) + \sin 2\pi f_0 t (N_0 \sin 4\pi f_0 \delta_{\max} T)\end{aligned} \tag{46}$$

Letting $\theta = 4\pi f_0 \delta_{\max} T$, the amplitude power of this normalized signal is:

$$a^2(\theta) = (N_0 + N_1 \cos\theta)^2 + N_1^2 \sin^2\theta \ . \tag{47}$$

The total number of pulses within the filter response time is $N = 1/B_h T = \overline{R}_p / B_h$, and $N_0 + N_1 = N$, so

$$a^2(\theta) = N^2 + 2NN_1(\cos\theta - 1) + N_1^2(2 - 2\cos\theta). \tag{48}$$

N_1 is a binomial random variable. Assuming "0" and "1" are equally likely, the variance of N_1 is $0.25N$ and its expected value is $0.5N$. Therefore,

$$E[N_1^2] = (E[N_1])^2 + \mathrm{var}(N_1) = 0.25(N^2 + N) \tag{49}$$

so

$$E[a^2(\theta)] = N^2 \cos\theta + 0.5(N^2 + N)(1 - \cos\theta), \tag{50}$$

which agrees with Figs. 3-5 for integer values of $f_0 T$. The average filter power output at integer values of $f_0 T$ (see Annex B) is $\overline{p}_y(f_0) = 2|G(f_0)|B_h^2 E[a^2(\theta)]$, assuming $|H_b(0) = 1|$. Therefore,

$$\overline{p}_y(f_0) = 2|G(f_0)|\left[\overline{R}_p^2 \cos 4\pi f_0 T\delta_{\max} + 0.5\left(\overline{R}_p^2 + \overline{R}_p B_h\right)(1 - \cos 4\pi f_0 T\delta_{\max})\right]. \tag{51}$$

Note that when $\cos 4\pi f_0 T\delta_{\max} = 1$, the magnitude of the spectral line is constant; the peak is equal to the average, so a "max hold" setting on the spectrum analyzer gives an accurate indication of the interference potential.

The main point here is that even an information-bearing pulsed signal may exhibit strong spectral lines unless specific measures are taken to prevent it. One such measure might be a pseudorandom dithering of the pulse positions. Of course, the transmitter and receiver would need to use the same dithering code, and must be synchronized with respect to the code phase.

Without such measures, it is apparent that a PPM signal could exhibit spectral lines, some of which can be as strong as those of an unmodulated pulse sequence with a rate equal to the average rate of the PPM sequence. However, with the PPM sequence, the average repetition rate cannot necessarily be inferred from the spacing between the spectral lines. For example, with $d_{\max} = 0.25$, the lines are $2/T$ apart. This spectrum could easily be confused with the spectrum of a constant-rate sequence with a rate that is double the actual average pulse rate of the PPM sequence.

4.6 Summary of Results

Letting $\Phi_g(f) = 2|G(f)|^2$ represent the energy spectral density of the pulse waveform $g(t)$, and f_0 be the center frequency of the narrowband filter, the power output of the filter response $y(t)$ to the pulse sequence $x(t) = \sum_{n=-\infty}^{\infty} g(t - T_n)$ is as follows for the indicated conditions; it has been assumed that $|H_b(0)| = 1$, as is the case for a spectrum analyzer.

- For $B_h > R_p$, the filter responds to each pulse individually and cannot resolve spectral lines. The peak envelope and average power are:

$$\begin{aligned} p_{y\,\text{peak}} &= \Phi_g(f_0)(kB_h)^2 \\ \overline{p_y} &= \Phi_g(f_0) R_p B_h \end{aligned} \qquad B_h > R_p \tag{52}$$

  The effective interference is taken here as the average over the duration of the filter response, which is:

$$p_y = \Phi_g(f_0) B_h^2 \tag{53}$$

- For $B_h << R_p$, and pulses repeated at a constant interval $T = 1/R_p$, the output power of a filter with $f_0 = mR_p$, where *m* is an integer, is a tone with envelope power:

$$p_y = \Phi_g(f_0) R_p^2 \qquad B_h << R_p \tag{54}$$

- For a pulse sequence with a random inter-pulse interval and average pulse rate $\overline{R}_P$, the filter output for $B_h << \overline{R}_P$ is a bandpass Gaussian process with average power:

$$\overline{p_y} = \Phi_g\left(f_0\right)\overline{R}_P B_h \tag{55}$$

- For a pulse-position modulated sequence with average rate $\overline{R}_P >> B_h$, there can be spectral lines with power $\Phi_g\left(f_0\right)\overline{R}_P^2$, but they generally will not be regularly spaced.

Figures 6 and 7 show the filter power output as a function of B_h and R_p, respectively, for regularly-spaced pulses.

The graphs in Figs. 6 and 7 are somewhat idealized, for illustrative purposes; the transitions will be smoother than shown at the break points where the bandwidth and pulse rate are equal. Experiments were conducted to verify these relationships; a pulse-modulated CW signal was fed into a spectrum analyzer. The generator output was +15 dBm, and the cable loss was about 1 dB. Figure 8 shows the power output for a 50-nanosecond rectangular pulse, with the pulse repetition frequency (PRF) varied. The resolution bandwidth setting on the analyzer was 100 kHz.

The energy spectral density in this case is $\Phi_g = p_g t^2 = -132$ dBmJ/Hz (decibels with respect to 1 millijoule per Hz). As expected, the output power is independent of the PRF for PRF < 100 kHz.

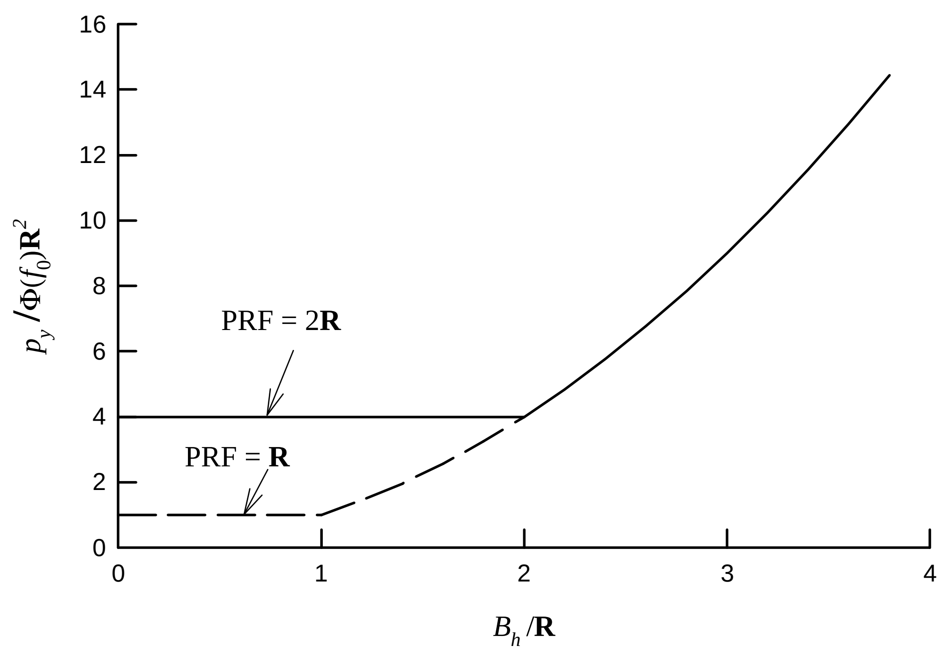


Figure 6: *Narrowband filter response to UWB signal (approximate for B_h near the PRF).*

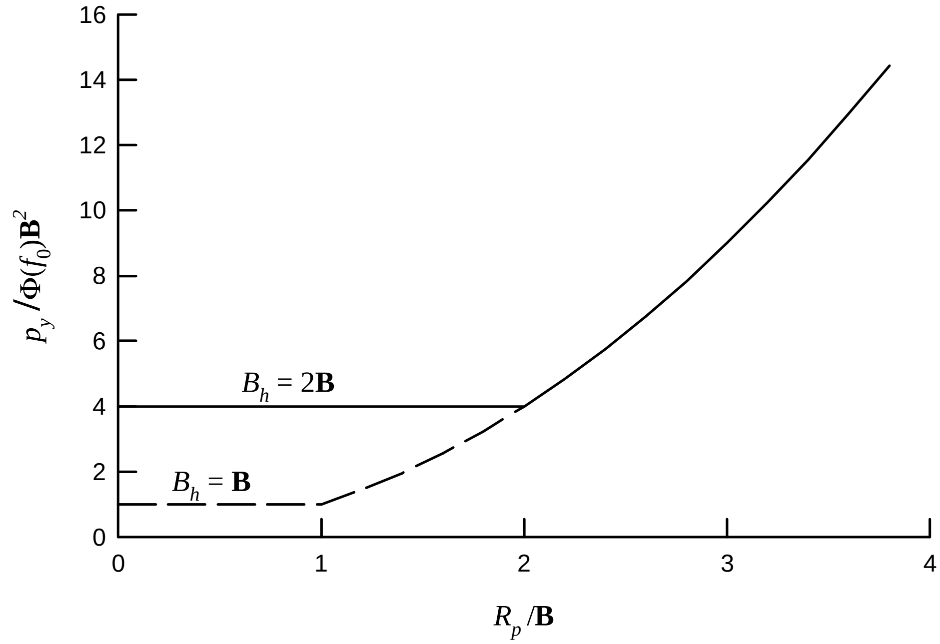


Figure 7: *Narrowband filter response to UWB signal (approximate for R_p near the filter bandwidth).*

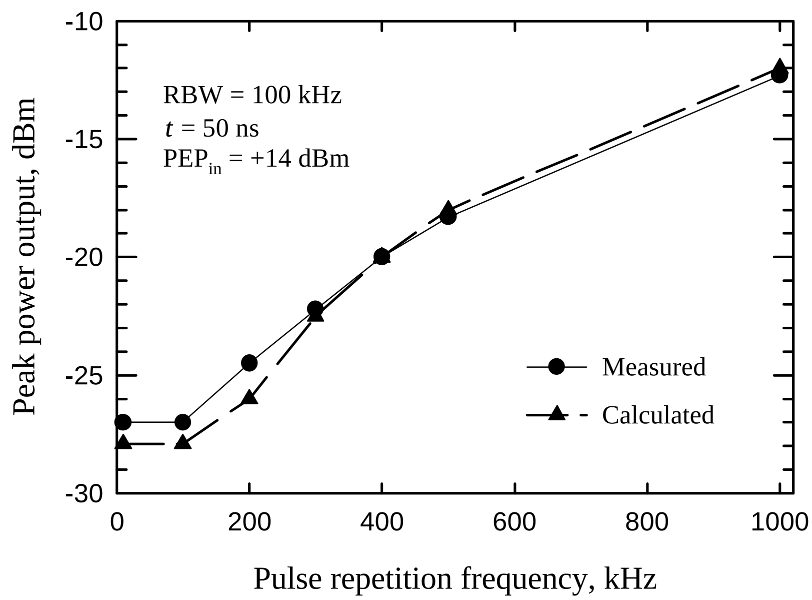


Figure 8: *Power output vs. PRF, measured vs. calculated.*

Figure 9 shows the power output vs. the resolution bandwidth, for a PRF of 100 kHz. The break-point seems to occur at a resolution bandwidth somewhat less than the PRF (in this case, 30 kHz).

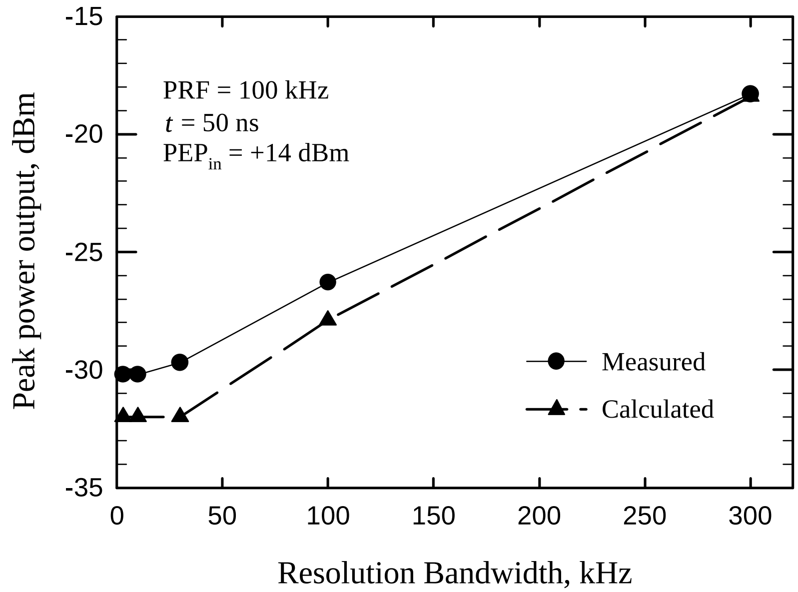


Figure 9: *Peak power output vs. RBW for PRF = 100 kHz.*

5. PROPAGATION PATH LOSS AND MULTIPATH

Detailed analysis of aggregate interference effects is beyond the scope of this paper, but as a prelude to such analysis, this section briefly discusses propagation effects which will need to be taken into account, and how they affect the interference from a UWB device that is experienced by a victim receiver.

The UWB transmitter and victim receiver generally will be spatially separated, so there will be a propagation path attenuation a_P. That is, if the UWB device transmits a pulse $g(t)$ with energy $E_{g,\mathrm{tx}}$, the received energy is $E_{g,\mathrm{rx}} = E_{g,\mathrm{tx}} / a_P$. It is $E_{g,\mathrm{rx}}$ which determines the strength of the interfering signal $I(t)$.

Often there will be multiple signal paths between a transmitter and receiver due to scattering and reflections. These paths generally are of different lengths, so multiple “copies” of the transmitted pulse arrive at the receiver, spread out in time. The “delay spread” is a measure of the span of arrival times for multipath components.

For a narrowband signal, the effect of multipath is to cause fading (fluctuations in the received signal power) as the transmitter or receiver moves around, or the positions of reflectors change. This fading is caused by changes in the phase relationships of the received multipath components, which leads to varying degrees of constructive and destructive interference. The result is the well-known “Rayleigh fading”, where the randomly fluctuating signal envelope has a Rayleigh distribution. If there is a line-ofsight path in addition to reflected/scattered paths, the distribution of the envelope tends to be Rician.

With a UWB signal, the pulse duration is likely to be significantly less than the delay spread, which tends to be on the order of one hundred or several hundred nanoseconds indoors, and microseconds or tens of microseconds outdoors. In that case, multiple distinct pulses of varying strengths and relative delays will arrive at the receiver. Figure 10 shows the UWB signal path from the transmitter, through the multipath channel, and through the narrowband filter $h(t)$ to produce the interference signal $I(t)$.

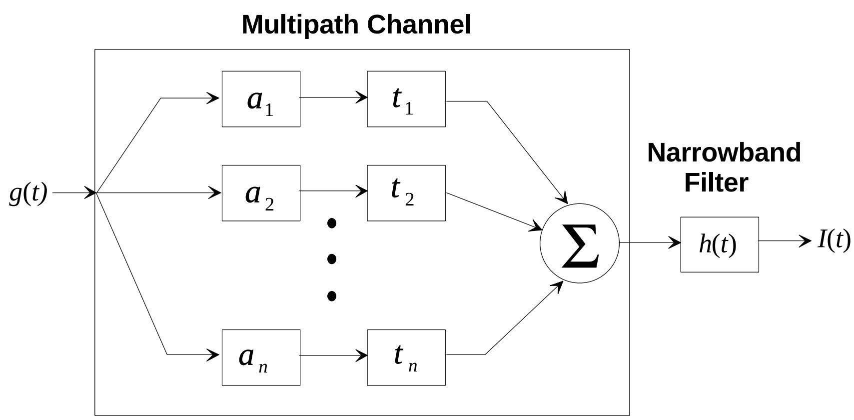


Figure 10: *Effect of the multipath channel on interference from a UWB pulse*

The $\{a_n\}$ and $\{t_n\}$ represent the attenuations and delays for the *n* signal paths.

For purposes of analyzing interference, the multipath channel and the narrowband filter may be interchanged as shown in Figure 11.

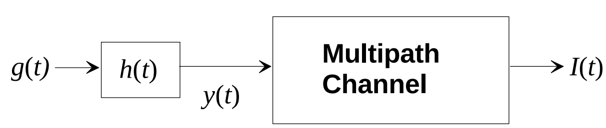


Figure 11: *Equivalent transfer function*

In Fig. 11, $y(t)$ represents the narrowband signal that would be transmitted, were the UWB signal passed through the narrowband filter $h(t)$ prior to transmission. The signal $y(t)$ then is subject to the usual effects of the multipath channel, including fading.

This approach simplifies analysis of aggregate interference from multiple UWB devices because the interference can be treated as the sum of multiple narrowband signals (which may be independently-faded) arriving from sources located random distances from the victim receiver. If the UWB devices transmit constant-rate pulse sequences with

$B_h < R_p$, the interference from each device will produce a tone at the receiver output. In general, the tones may be of slightly different frequencies because the repetition rates of the different devices will likely be different due to clock rate tolerance. In the victim receiver, the tones therefore will add non-coherently (power addition).

If the $B_h > R_p$, then the victim receiver will see multiple pulse sequences that are interleaved or superimposed on one another. In the latter case, the component pulses add coherently, because the frequency of the filter output is determined by the filter itself, not the input pulse. The magnitude of the composite pulse will depend on amplitudes and phase relationships of the components.

Clearly, the effect of aggregate interference will depend on a number of factors, including the characteristics of the UWB devices and victim receivers, as well as the density of UWB devices (average number of simultaneously active UWB devices per square km). Analysis of aggregate interference from UWB devices is a potential area for further work. It has been shown in this section how the results developed in this paper, which yield the signal $y(t)$, may be applied directly to the problem of analyzing aggregate interference from multiple UWB devices.

ANNEX A

Impulse Response of an *n*-Pole Filter

If the baseband-equivalent transfer function of the spectrum analyzer resolution filter is

$$H_b(f) = \frac{1}{(1 + j2p\ f/a)^n} \tag{A-1}$$

then its impulse response is

$$h_b(t) = \frac{1}{(n-1)!} a^n t^{n-1} e^{-a\,t}, \quad t \geq 0 \tag{A-2}$$

Since $|H_b(0)| = 1$, the effective noise bandwidth of the filter is

$$B_h = \int_{-\infty}^{\infty} |H_b(f)|^2 df = \frac{a}{2p} \int_{-\infty}^{\infty} \frac{dx}{(1+x^2)^n} = \frac{a}{2p} \frac{\Gamma\left(\frac{1}{2}\right)\Gamma\left(n - \frac{1}{2}\right)}{(n-1)!} \tag{A-3}$$

where $\Gamma(\cdot)$ is the Gamma function. Since $\Gamma(n+1) = n\Gamma(n)$ and $\Gamma\left(\frac{1}{2}\right) = \sqrt{p}$,
$\Gamma\left(\frac{1}{2}\right)\Gamma\left(n - \frac{1}{2}\right) = p\left[\frac{1}{2}\cdot\frac{3}{2}\cdots\left(n - \frac{3}{2}\right)\right]$.

The time at which $h_b(t)$ achieves its maximum response can be found by setting $dh_b(t)/dt = 0$ and solving for *t*. The result is:

$$t_m = \frac{n-1}{a}, \tag{A-4}$$

which might be regarded as the filter "rise time". The maximum response is:

$$h_{b\max}(t) = h_b(t_m) = \frac{a\,(n-1)^{n-1}}{(n-1)!} e^{1-n} \tag{A-5}$$

The power output of the filter measured using a "max hold" setting is $h_b^2(t_m)$. Table A-1 below shows a, t_m, and $h_b(t_m)$ for n = 2, 3, and 4 poles. The values for a can be checked by substituting in the impulse response formula and verifying that $\int_0^\infty h_b^2(t)dt = B_h$.

Table A-1: *Maximum response of an n-pole filter*

n	a	t_m	$h_b(t_m)$
2	$4B_h$	$1/4B_h$	$1.47B_h$
3	$16B_h/3$	$3/8B_h$	$1.44B_h$
4	$32B_h/5$	$15/32B_h$	$1.43B_h$

Note that in all cases, $h_b(t_m) \cong 1.45B$.

The nominal "resolution bandwidth" is typically the 3-dB bandwidth of the filter.[1] Table A-2 shows $B_h/B_{3\text{dB}}$, the ratio of the noise bandwidth to the 3-dB bandwidth, as well as well as $h_b(t_m)$ as a function of $B_{3\text{dB}}$.

Table A-2: *Ratios of noise bandwidth and maximum response to 3-dB bandwidth for n-pole resolution filter.*

n	$B_h/B_{3\text{dB}}$	$h_b(t_m)/B_{3\text{dB}}$
2	1.220	1.79
3	1.155	1.66
4	1.128	1.61

The quantity $h_b(t_m)$ is the so-called "impulse bandwidth" B_{imp} that is used in computing pulse desensitization.[2] The manual for the HP 8560 E-series analyzers gives B_{imp} as 1.6 times the nominal resolution bandwidth.

Letting $x = a/B_h$ (which from table 1 is a function of n), $h_b(t)$ can be expressed as:

$$h_b(t) = \frac{B_h x^n}{(n-1)!}(B_h t)^{n-1} e^{-xB_h t} \tag{A-6}$$

and a normalized version of $h_b(t)$ is

[1] Hewlett-Packard, Application Note 1303, "Spectrum Analyzer Measurements and Noise," page 9.
[2] Hewlett-Packard, Application Note 150-2, "Spectrum Analysis – Pulsed RF," November 1971.

$$h_n(t) = \frac{h_b(t/B_h)}{B_h} = \frac{x^n}{(n-1)!} t^{n-1} e^{-xt}. \qquad \text{(A-7)}$$

Hence, $h_b(t) = B_h \cdot h_n(B_h t)$. Note that $\int_0^\infty h_n(t)dt = \int_0^\infty h_b(t)dt = H_b(0)$. Figure A-1 shows $h_b(t)/B_h$ vs. $B_h t$.

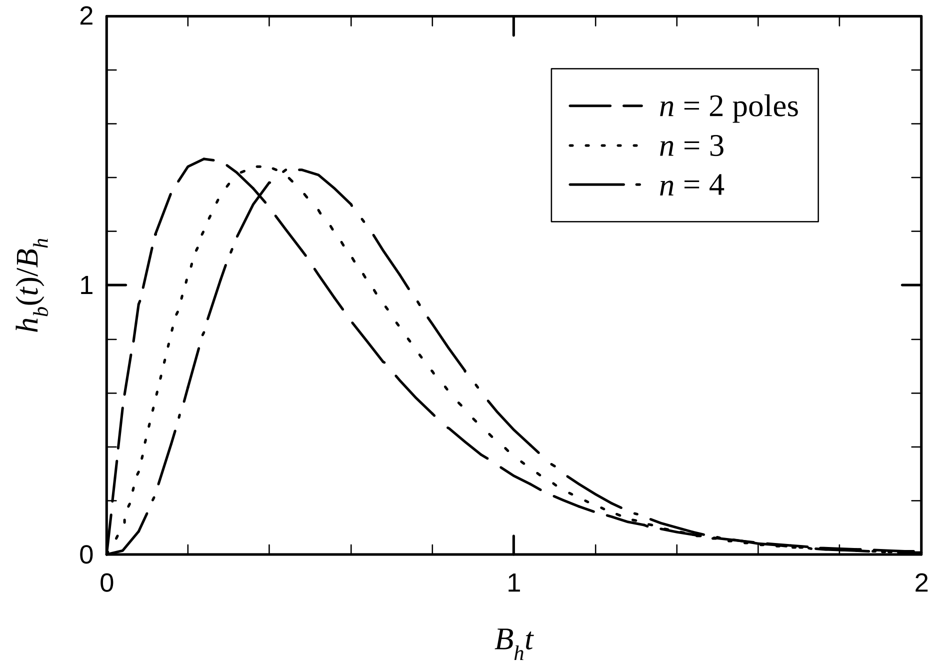


Figure A-1: *Normalized n-pole filter impulse response.*

ANNEX B

Simulating the Narrowband Filter Response to a UWB Signal

The in-phase and quadrature components of the filter output are:

$$C_I = \sum_{k=0}^{L} h_b(t - T_k)\cos 2pf_0T_k$$
$$C_Q = \sum_{k=0}^{L} h_b(t - T_k)\sin 2pf_0T_k \qquad \text{(B-1)}$$

If pulse-position modulation (PPM) is applied to the pulse sequence, $T_k = kT + d_kT$, where d_k is the time-shift of the k^{th} pulse as a fraction of the nominal pulse interval *T*; d_k is a random variable with properties determined by the specific modulation characteristics.

The number of components *L* must be chosen so that the sum is taken over the duration of the impulse response, which from Fig. A-1 is $2/B_h$. Hence, a reasonable choice for *L* is $L = \text{int}(2/B_hT)$. Since the case of interest here is $B_hT << 1$, $1/B_hT$ (which will be a parameter for the simulation) can be constrained to be an integer, and

$$L = \frac{2}{B_hT} \qquad \text{(B-2)}$$

The filter is causal ($h(t) = 0$ for $t < 0$), so *t* must satisfy $t \geq LT + d_LT$. This requirement can be met by setting *t* to:

$$t = T(L + d_{\max}) \qquad \text{(B-3)}$$

To normalize the simulation results, the normalized version of the baseband impulse response $h_n(t) = h_b(t/B_h)/B_h$ can be used; see eq. (A-7). Defining

$$T_n = B_hT = 2/L \qquad \text{(B-4)}$$

$$t_n = B_ht = T_n(L + d_{\max}) = 2 + 2d_{\max}/L, \qquad \text{(B-5)}$$

the amplitude of the filter response to the k^{th} pulse becomes:

$$h_b(t - T_k) = B_hh_n\left[t_n - \frac{2}{L}(k + d_k)\right] \qquad \text{(B-6)}$$

and the normalized in-phase and quadrature components are:

$$C_I' = C_I / B_h = \sum_{k=0}^{L} h_n \left[t_n - \frac{2}{L}(k + d_k) \right] \cos 2pf_0 T (k + d_k) \qquad \text{(B-7)}$$

$$C_Q' = C_Q / B_h = \sum_{k=0}^{L} h_n \left[t_n - \frac{2}{L}(k + d_k) \right] \sin 2pf_0 T (k + d_k). \qquad \text{(B-8)}$$

The filter output envelope power is

$$p_y = 2 \left| G(f_0) \right|^2 B_h^2 \left(C_I'^2 + C_Q'^2 \right) = \Phi_g(f_0) B_h^2 \left(C_I'^2 + C_Q'^2 \right). \qquad \text{(B-9)}$$

Given $B_h T$, $f_0 T$, and a function to generate the $\{d_k\}$, the simulation generates a large number of samples of C_I' and C_Q' , and accumulates their distributions as well as the average power. Output format can be varied according to the purpose. The distributions can be computed for a single value of $f_0 T$, or average power can be computed as a function of $f_0 T$ over some desired range. In the latter case, a plot of the output would be comparable to the display of a spectrum analyzer; that is, the power output of the resolution filter as a function of the filter center frequency.

The simulation results can be checked several ways. If $f_0 T$ is an integer and there is no modulation, $C_Q' = 1/B_h T$ for $B_h T << 1$. This is because

$$\sum_{k=0}^{L} h_n (k B_h T) \cong \frac{1}{B_h T} \int_0^{\infty} h_n(t) dt = \frac{H_b(0)}{B_h T} \qquad \text{(B-10)}$$

and it is assumed here that $H_b(0) = 1$. Therefore, the power in each spectral line of the simulation output, for a periodic pulse sequence, should be $(1/B_h T)^2$.

A second test is to vary $f_0 T$ over some small range about an integer value, with a periodic sequence as the input, and plot the filter frequency response. Figure B-1 shows an example for $1/B_h T = 20$. From Annex A, the response of a 4-pole filter at a frequency Δf from its center frequency should be:

$$\left| H(f_0 \pm \Delta f) \right|^2 = \frac{1}{\left[1 + \left(\frac{10p}{32} \cdot \frac{\Delta f}{B_h} \right)^2 \right]^4} \qquad \text{(B-11)}$$

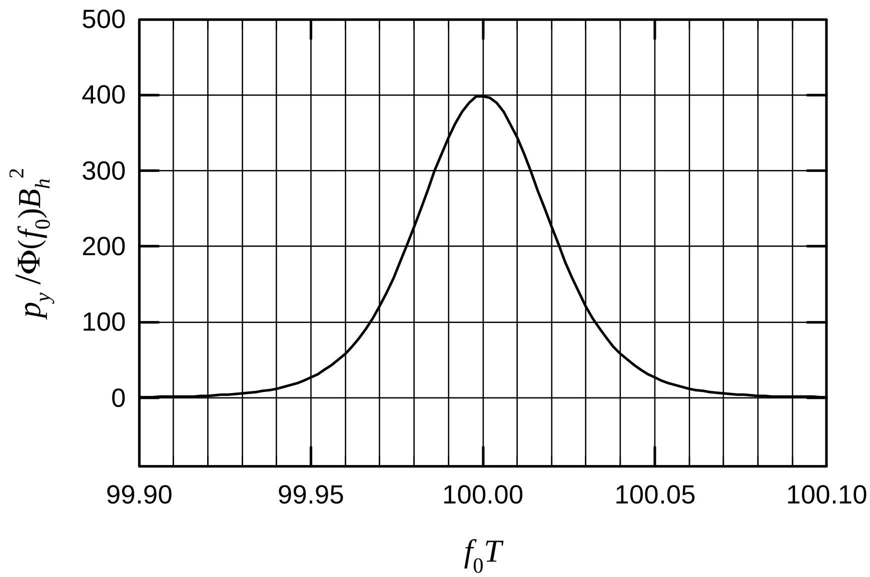


Figure B-1: *Simulation output for 1/B_hT = 20, a periodic sequence (no modulation), and a 4-pole filter.*

The simulation result shown in Fig. B-1 is in agreement with (B-11). Also, the maximum power is $(1/B_hT)^2$, as it should be.

A third test is to vary the pulse positions randomly and plot the distribution of C_I' and/or C_Q'. Figure B-2 shows the distribution of C_I' from the simulation for a uniform distribution of pulse positions between 0.9*T* and 1.1*T* (i.e., $d_{\max} = 0.1$), and $1/B_hT = 20$. A Gaussian scale is used, meaning that a Gaussian distribution appears as a straight line. It can be seen that the distribution is zero-mean Gaussian. The standard deviation is 3.15 (the 90th and 10th percentile points are $1.29s$ above and below the median), so the variance is 9.9. This is as it should be. From (44) $E\left[C_I^2\right] = E\left[C_Q^2\right] = \overline{R}_P B_h/2$ (for $\left|H_b(0)\right|^2 = 1$), and $\overline{R}_P = 1/T$. Since $C_I' = C_I/B_h$ and $C_Q' = C_Q/B_h$,

$$E[C_I'^2] = E[C_Q'^2] = 1/2B_hT \qquad \text{(random PPM)} \qquad \text{(B-12)}$$

Thus, in this case $E[C_I'^2] = E[C_Q'^2] = 10$.

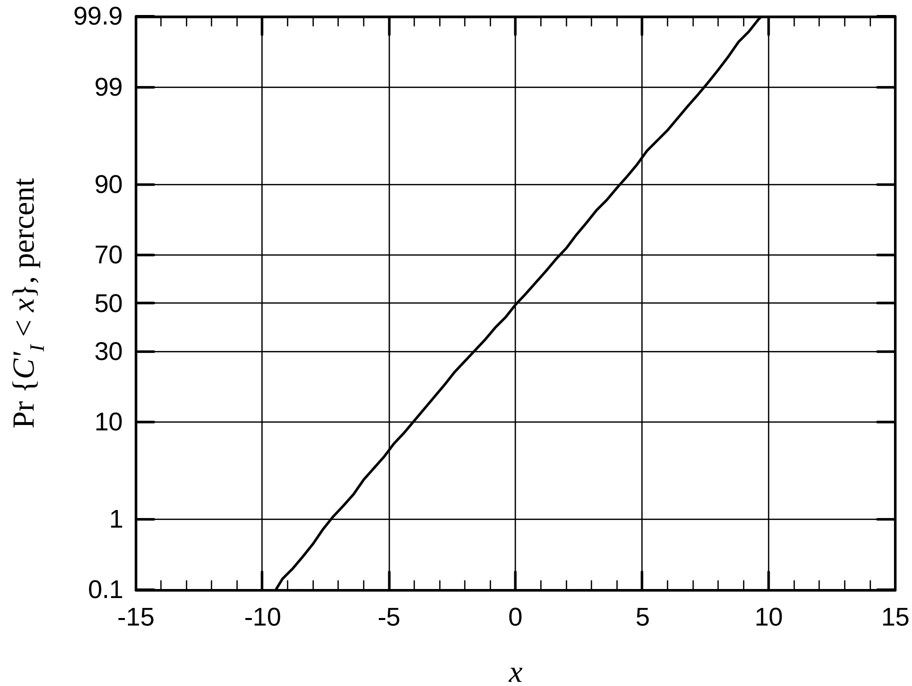


Figure B-2: *Distribution of C'_I for $1/B_hT = 20$ and randomly-shifted pulses with $d_{\max} = 0.1$.*

ATTACHMENT 2

Comparison of Some Potential Emissions Specifications for Ultra-WideBand Devices

Donald Johnson
December 7, 1998

Introduction and Summary

The FCC Notice of Inquiry (NOI) regarding Ultra-WideBand (UWB) transmission systems poses a number of questions concerning the proper specifications and measurement procedures for controlling the emissions in the broadcast and restricted bands. This paper addresses these questions by describing a range of potential emission types and comparing their relative interference potential. Section 15.309 and 15.35 require that unwanted emissions at frequencies above 1000 MHz be measured with a 1 MHz resolution bandwidth filter and the limits are expressed as a voltage average with a further limitation on the peak envelope power. This means of specification and control of unwanted emissions has the advantage of being easily measurable, but it has drawbacks. In some cases it understates the effect the emission will have on a potential victim receiver and in other cases it overstates the effect. This effect is compared quantitatively for the range of emission types.

It is shown that the current rules permit high relative levels of interference to narrow bandwidth victim receivers for some emission types and that they permit relatively high levels of interference to wide bandwidth receivers for other emission types. At the same time, the rules are excessively strict for some emission types.

The emission specifications for UWB devices within the restricted or otherwise controlled bands and the out-of-band emissions of other wide bandwidth packet transmission devices need to be improved to reflect the interference effect of the emissions more evenly and directly. The general principle should be that the specified parameters should relate directly to the potential for interference in victim receivers**.** An enhanced specification based primarily on burst PSD and utilizing other measurement bandwidths is required to address the emissions of UWB devices and other wide bandwidth, packet transmission devices.

Power Spectral Density and Peak Envelope Power

The NOI, in paragraph 12, asks whether the Power Spectral Density (PSD) should be the basis for the specification or whether different limits should be applied. WINForum has advocated burst PSD[1] as the basis for specification of out-of-band emissions for the U-NII band. However, there are instances where the peak interference power may be the proper specification parameter. In some cases the peak power is a predictable multiple of

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[1] The burst power spectral density (burst PSD) as used here is the parameter of definition (f) in §15.403 Definitions.

> 15.403 (f) *Power Spectral Density*. The power spectral density is the total energy output per unit bandwidth from a pulse or sequence of pulses for which the transmit power is at its peak or maximum level, divided by the total duration of the pulses. This total time does not include the time between pulses during which the transmit power is off or below its maximum level.

the burst PSD and, in such cases, either can be used with proper account taken of the multiplying factor.

When the emission is the out-of-band product of a wide bandwidth, relatively long duration transmission burst, the short-term average PSD is the appropriate parameter to control this. In this case the signal is very noise-like and the peak power is related only statistically to the burst PSD. In some instances it is necessary to control the local peak PSD measured over intervals of a few reciprocal bandwidths of the measurement filter or victim receiver rather than the average over the complete burst[2]. The use of a video filter of narrower bandwidth than the resolution bandwidth is one method of controlling this.

The emission power level is normally specified as the power level in a particular bandwidth. This is an approximation to the PSD. The bandwidth with which the burst PSD is approximated in this manner should be equal to or lower than the expected bandwidth of the victim receivers in most cases. The 1 MHz bandwidth currently used in Part 15 is appropriate only if the expected victim bandwidths exceed this value.

Some Emission Types

This section defines some types of emissions that represent the range of possible types which UWB and other wide bandwidth devices might generate.

It is shown in attachment 1 that the emissions of UWB devices may consist of anything from discrete spectral lines for a periodic waveform to a continuous, relatively flat spectrum for random waveforms. The UWB emission specification should take this into account. Thus, the specification should control both emissions consisting of multiple very narrow line spectra and those consisting of very wide bandwidth regions of relatively continuous spectra.

Attachment 1 also shows that emissions created by periodic UWB pulses which consist of multiple discrete lines within the victim or measurement bandwidth (the repetition rate is less than the victim bandwidth) have peak and burst PSD levels proportional to the square of the measurement or victim bandwidth. These are in contrast to the normal emission types of digital communication transmitters, thus they need special consideration.

A worst type of interfering emission signal which passes the current specification and measurement is a narrow tone signal that bursts on and off with a duty cycle of under 10% over any 100 ms interval[3]. This is the worst case signal with respect to narrow

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2 Attachment B to the WINForum Petition for Reconsideration in ET Docket No. 96-102, Amendment of the Commission's Rules to Provide for Operation of Unlicensed NII Devices in the 5 GHz Frequency Range addresses an instance where the PSD needs to be controlled over a shorter interval than the complete burst length. Transmission bursts with short duration, high power levels during longer transmission bursts with lower average power are studied. This type of transmission burst leads to a non-stationary process in which the power level measured over a short segment of a transmission consisting of a few reciprocal bandwidth time intervals is high relative to the average of the burst. A means of computing the relative effects of these short duration high levels during a burst is developed and it is shown that such transmissions are detectable with common measurement instruments.

3 Pulsed operation and the 100 ms interval are specified in §15.35 (c). When the duty cycle is under 10% the peak power is the limiting parameter.

bandwidth victim receivers. The Equivalent Isotropic Radiated Power (EIRP) of such a signal can be as high as -21 dBm and meet the current requirement[4]. For bandwidths of 1 MHz and greater, the permissible peak power level is -21 dBm per MHz. This power level is the same for narrower victim bandwidths, that is, the PSD for narrower bandwidths can be inversely proportional to bandwidth. Since the background noise level is lower for narrower bandwidths, the narrower bandwidth receiver is likely to be sensitive to a lower level.

An UWB device that sends sequences of pulses at a high constant repetition rate for a short duration (a few milliseconds) with repetitions of the sequences over short intervals might create such a signal, for instance. Such sequences may be very seldom generated. However, in the restricted bands, where safety of life systems may be located, it is necessary to plan for the worst eventuality. Thus, the worst case sequence should be anticipated.

UWB devices that send information carrying signals may generate wide bandwidth continuous spectra signals (greater than 1 MHz emission bandwidths). Consider, for example, an information bearing signal with a baseband bandwidth of B_i ($2B_i$ > 1 MHz) which modulates a repetitive RF pulse of repetition rate R_p, in which $R_p >> 2B_i$. The Fourier transform of such a signal consists of multiple repetitions of an emission mask of bandwidth $2B_i$ separated in frequency by units of R_p. The statistics of such a signal when measured with a narrow bandwidth filter (bandwidth << B_i) will be much the same as the out-of-band signals generated by communication transmitters of bandwidth greater than the measurement bandwidth. WINForum described this type of emission in an earlier paper[5].

In the following, five general emission types are compared with respect to the burst PSD of the signal level which just meets the current restricted band specifications for signals of frequency above 1 GHz. These represent the range of emissions that can be expected from UWB devices as well as some that can be expected from more conventional systems such as wide bandwidth U-NII systems.

1. The worst case signal described above. Very narrow bandwidth, constant envelope with burst duty cycle less than 10%.
2. Wide bandwidth bursts (B_i >>1 MHz) with duty cycle less than 10%
3. Very narrow bandwidth, constant envelope signal with burst duty cycle of 100%.
4. Wide bandwidth bursts (B_i >>1 MHz) with burst duty cycle of 100%.
5. A periodic pulsed signal with repetition rate lower than the measurement bandwidth ($B_m >> R_p$).

4 The requirement is that the field strength of the radiated field 3 meters from the transmitter in the direction of maximum radiation be no higher than 5 millivolts/meter (20 dB above 500 microvolts/meter). This is equivalent to an EIRP value of –21.2 dBm.

5 Attachment A to the WINForum Petition for Reconsideration in ET Docket No. 96-102, Amendment of the Commission's Rules to Provide for Operation of Unlicensed NII Devices in the 5 GHz Frequency Range, Wideband Emissions Through A Narrowband Filter and the Implications on Measurement of Power Spectral Density Using a Spectrum Analyzer, March, 1997.

Comparison of Four Signal Types

Signal types 1 through 4 are candidates for a specification based on burst PSD if the victim bandwidth is less than 1 MHz. The 4 types have higher interference potential in receivers of narrow bandwidth and are best characterized using a measurement bandwidth less than or equal to the potential victim receiver bandwidth.

The current rules specify and control the peak envelope power of the low duty cycle signals (types 1 and 2). Measurement of the peak of a wide bandwidth, low duty cycle signal (type 2) presents a problem because of the statistical nature of the envelope variations. This signal is noise like and the peaks may be widely separated in time. Further, the peak values depend upon the digital sequences used for testing transmitters. Burst PSD is a better parameter to specify for these noise-like signals.

The following table compares the burst PSD these 4 signal types create in a potential victim receiver at 1 km separation in a representative environment.

The following conditions apply to the table:

Propagation model – free space distance attenuation with 14 dB excess attenuation.
Receiver noise figure: 6 dB
Wide bandwidth signal peak/average at 1 MHz bandwidth[6] : 6 dB
Noise level –108 dBm =10 log kTB + 6 dB

Let

P_a = the peak to average power ratio. P_a = 0 for narrow bandwidth signals.
L = the interference burst PSD-to-noise power ratio (in deciBels) at 1 km separation.

The equations for the table values are

$L_{1MHz} = -21.2 - P_a - 14dB - 20Log\left(1000/3\right) + 108 = 22.3 - P_a$ dB for a duty cycle under 10 % and

$L_{1MHz} = -41.2 - 14dB - 20Log\left(1000/3\right) + 108 = 2.30$ dB for a full duty cycle signal.

It is assumed that the voltage average is the same as the power average for the full duty cycle signal.

The interference ratio of the narrow bandwidth signals (types 1 and 3) increase in inverse proportion to the victim bandwidth for bandwidths less than 1 MHz since the actual emission power remains the same at lower bandwidths. On the other hand, the interference ratio of the wideband signals tends to remain constant because of the relatively smooth PSD over the 1 MHz band.

[6] This type of signal has a randomly varying envelope and approaches a gaussian distribution as the measurement bandwidth is reduced relative to the emission bandwidth of the signal. A measurement on a QPSK signal with a 10:1 emission bandwidth to measurement bandwidth ratio showed a 6 dB peak-to-average ratio. Thus, the conditions of the table represent about a 10 MHz signal bandwidth.

	Interference-to-Noise at 1 Km (dB)		
Emission Signal Description	**Victim Bandwidth = 1 MHz**	**Victim Bandwidth = 100 kHz**	**Victim Bandwidth = 10 kHz**
1. Narrow bandwidth, constant envelope, duty cycle< 10%	22.3	32.3	42.3
2. Wide bandwidth bursts, duty cycle < 10%, P_a = 6 dB	16.3	≈16.3 (same as 1 MHz)	≈16.3 (same as 1 MHz)
3. Narrow bandwidth, constant amplitude, duty cycle = 100%	2.3	12.3	22.3
4. Wide bandwidth, duty cycle = 100%, P_a = 6 dB	2.3	≈2.3 (same as 1 MHz)	≈2.3 (same as 1 MHz)

Table 1: Interference-to-Noise Levels with the Maximum Emission Level Permitted by the Current Rules for the Four Emission Types.

Table 1 compares the interference-to-noise levels for the four emission types at a separation distance of 1 km. The comparison parameter is mean power level to noise power ratio while the signals are on. The wide bandwidth signals have envelope variations so that the peak envelope power is greater than the mean power. Nevertheless, in most instances, it is the mean power level that determines the interference effect. In the one case where a single device is the only interferer, and the interference level exceeds the noise level, the equivalent level relative to the interference effect is slightly higher than the mean. However, in most cases the mean is a better measure and the signals of the table have nearly equal effect.

The limitation of the current emission specification and measurement technique is best illustrated by the case of the lowest victim bandwidth. Here the worst case signal creates an interference-to-noise ratio nearly 40 dB higher than does the wide bandwidth, full duty cycle (type 4) signal. This problem could be prevented with a specification on burst PSD with a bandwidth resolution as low as that of the expected victim receivers.

Wideband high-speed digital communication transmitters tend to generate wide bandwidth out-of-band signals. A specification based on burst PSD with a low bandwidth resolution and a specified level consistent with the level currently required by the rules would be favorable to such intentional transmitters as well provide more protection to restricted band receivers with bandwidths narrower than 1 MHz.

Periodic Pulsed Signal with Low Repetition Rate

As shown in attachment 1, the peak power level and burst PSD of a periodic, pulsed signal with repetition rate lower than the measurement or victim bandwidth (type 5 signal above) is proportional to the square of the bandwidth. At low relative repetition rates the filter response duration is less than the pulse spacing and the waveform is a sequence of individual responses of the measurement filter to the short duration wideband signals.

The measurement bandwidth for frequencies above 1000 MHz is now 1 MHz, thus the UWB periodic signals with repetition rates below about 1 million pps are of this type if the current specification and measurement procedure is used.

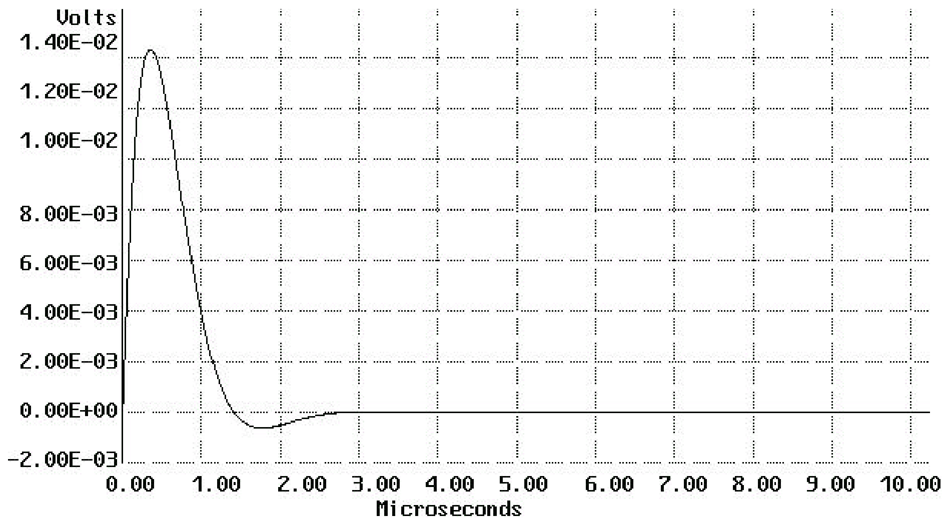

Figure 1: Impulse Response of a Two Pole Butterworth Low Pass Filter
Peak = 14.3 mv
RMS over 1.4 microseconds = 8.89 mv
Peak to RMS = 4.1 dB

Figure 1 is an approximation of the impulse response of a measurement filter with a resolution bandwidth of 1 MHz. The figure shows the impulse response of a 500 kHz bandwidth low pass filter that is approximately the baseband equivalent impulse response of a 1 MHz resolution bandwidth filter. An UWB device transmitting wideband pulses at 700 pulses per second would create a waveform in the measurement filter consisting of a sequence of pulses with this envelope shape separated by 1.4 microseconds. Such a signal has a peak to average burst power ratio of about 4 dB. That is, the burst PSD is about 4 dB below the peak power. Periodic signals with this repetition rate and less fit the definition of type 5 signals with respect to the current measurement bandwidth of the rules. Thus, the peak power is about 4 dB above the burst PSD for type 5 signals.

Either the peak or the burst PSD can be used as the specification parameter. However, the peak level is easiest to measure over the wide frequency range of UWB devices and may be a better parameter to use for specifying the emission, since it is so directly related to the burst PSD.

Table 2 compares the interference effect of a signal just meeting the current emission specifications on a repetitive wideband pulse at various repetition rates and at three victim bandwidths.

The interference created by this signal in a victim receiver with a bandwidth greater than the repetition rate consists of a regular pulse with high peak power. The proper comparison point at other bandwidths when the current Part 15 emission rules are applied is the burst PSD or peak power divided by the bandwidth, as this represents the peak power to thermal noise ratio. Table 2 shows the approximate values of the burst PSD to bandwidth ratio for repetitive wideband pulse sequences all of which have the same peak envelope power at a 1 MHz measurement bandwidth. The table thus serves to compare the effects when the current rules are used.

	Relative Peak Power per Unit Bandwidth (dB)		
Repetition Rate	Victim Bandwidth = 100 kHz	Victim Bandwidth = 1 MHz	Victim Bandwidth = 10 MHz
7 Kpps	-10	Reference	+10
70 Kpps	-10	Reference	+10
700 Kpps	+4	Reference	+10
7 Mpps	+10	Reference	-4
70 Mpps	+10	Reference	-10

Table 2: Peak Power per Unit Bandwidth for Constant Peak Envelope Power at 1 MHz
This applies when the level is controlled by the current Part 15 specification

The signal is of type 5 in the first row of the table at all bandwidths and is a type 3 signal (single spectral line) in the last row at all bandwidths. The table shows that the current emission specification over-protects low bandwidth victim receivers from low repetition rate pulse sequences. It also shows that the current rules understate the interference potential for this type of interference in receivers of bandwidth greater than 1 MHz. The opposite effect is demonstrated in the last rows of the table.

A high ratio of victim bandwidth to repetition rate works to some extent to reduce the effect of the high peak power bursts in digital information receivers if the effect is taken into account in the receiver design. For example, the information rate in the 10 MHz receiver might be 10 Mb/s, in which case the 7000 pulse per second interferer would create interference about every 1400 bits. Burst error correction would greatly alleviate this interference. However, this has little effect on receivers that are already designed and thus it does not reduce the peak power effect in all receivers.

The aggregation effect of a type 5 signal is also lower if the repetition rate is low. The probability of two interfering UWB devices having overlapping pulses is proportional to the repetition rate divided by the victim bandwidth for such signals, thus there is a lower instance in which signal peaks add. The aggregation effect requires further study.

Conclusions

The current Part 15 emission specification is deficient for UWB devices in that the specified parameters do not relate directly to the interference effect of some types of emissions that are likely to be generated by such devices.

An UWB device that transmits sequences of pulses at a rate lower than the measurement or victim receiver bandwidth creates regular peak power levels of potentially high magnitude. The PSD of such emissions is proportional to the bandwidth of the victim receiver. Thus, the rules under protect receivers of bandwidth greater than the current 1 MHz measurement bandwidth and over protect narrower bandwidth receivers.

On the other hand, the current rules do not sufficiently protect narrow bandwidth victim receivers from other common types of emissions.

Emission rules based on burst power spectral density are needed to control UWB devices. Such rules can be crafted to more consistently control the interference potential of all emission types. They will better limit the interference potential than the current rules for higher interfering emission types and should relax the requirements for less interfering emission types.

A power spectral density limit appropriately controls all types of unwanted emissions but a peak power limit is appropriate in some circumstances.